

DESCRIPTION

WIRELESS COMMUNICATIONS SYSTEM AND WIRELESS DIGITAL RECEIVER FOR
USE THEREIN

5

TECHNICAL FIELD

The present invention relates to a wireless communications system and a wireless digital receiver for use therein, and more particularly to a wireless communications system using an FDD (Frequency Division Duplex) architecture and a wireless digital receiver for use therein.

BACKGROUND ART

One of the conventional wireless communications systems using the FDD architecture is a wireless communications system using a DSRC (Dedicated Short Range Communications) architecture (hereinafter referred to as a "DSRC system").

The DSRC system standard specifies that when transmitting a first wireless signal from a first wireless communications device provided on the road (hereinafter referred to as a "base station") to a second wireless communications device provided in a vehicle (hereinafter referred to as a "mobile station") (hereinafter such a transmission will be referred to as a "downlink"), one of 5775 [MHz], 5780 [MHz], 5785 [MHz], 5790 [MHz], 5795 [MHz], 5800 [MHz] and 5805 [MHz] is used as the center frequency.

The DSRC system standard also specifies that when transmitting a second wireless signal from the mobile station to the base station (hereinafter such a transmission will be referred to as an "uplink"), a center frequency that is away from that used for the downlink by 40.000 [MHz] is used. Specifically, if 5775 [MHz] is used as the center frequency for the downlink, 5815 [MHz] is used as the center frequency for the uplink. Similarly, if 5780 [MHz] is used for the downlink, 5820 [MHz] is used for the uplink. If 5785 [MHz] is used for the downlink, 5825 [MHz] is used for the uplink. If 5790 [MHz] is used for the downlink, 5830 [MHz] is used for the uplink. If 5795 [MHz] is used for the downlink, 5835 [MHz] is used for the uplink. If 5800 [MHz] is used for the downlink, 5840 [MHz] is used for the uplink. If 5805 [MHz] is used for the downlink, 5845 [MHz] is used for the uplink.

In a section on the technical requirements for wireless equipment in DSRC system standard specifications, standards for image response are specified only for the base station.

Where a demodulation process is performed by a digital signal processing circuit, in order to convert a received modulated high-frequency signal to a frequency suitable for the digital signal processing circuit, a frequency conversion circuit for downconverting the modulated high-frequency signal needs to be provided preceding the digital signal processing circuit.

In view of the technical requirements for wireless equipment, it is preferred that a frequency conversion circuit employing a

LOW-IF architecture, for example, is used in the base station. This is because it is possible with the LOW-IF architecture to remove an image disturbing signal without using an image suppression filter in a high-frequency part, as described in
5 Non-Patent Document 1 (J. Crols and Michiel S. J. Steyaert, "Low-IF Topologies for High-Performance Analog Front Ends of Fully Integrated Receivers", IEEE TRANSACTIONS ON CIRCUITS AND SYSTEMS II: ANALOG AND DIGITAL SIGNAL PROCESSING, VOL. 45, NO. 3, March 1998).

10 As described in Non-Patent Document 1, with the LOW-IF architecture, the center frequency of a received modulated high-frequency signal is downconverted to a frequency that is about a few times as great as the signal bandwidth of the modulated high-frequency signal. Then, the downconverted signal is
15 directly sampled by a sampler and demodulated by a digital signal processing circuit. The LOW-IF architecture is advantageous in that it offers better reception characteristics and high degrees of integration.

~~For the mobile station, however, no image response standard~~
20 is specified. Therefore, it is possible to use a frequency converter in which a local oscillator is shared by the transmitter and the receiver. Thus, with the mobile station, a single-conversion architecture can be employed. Therefore, the mobile station can be provided at a low cost.

25 As described above, where a frequency converter employing

the LOW-IF architecture is used in the base station, a frequency-converted signal is converted into a signal having a frequency that is about a few times as great as the signal bandwidth of the received modulated high-frequency signal.

5 Where the mobile station uses a frequency converter employing the single-conversion architecture in which a local oscillator is shared by the transmitter and the receiver, a frequency-converted signal is converted into a signal having a frequency that is equal to the difference between the uplink and
10 the downlink. Typically, the frequencies are different from each other. This is illustrated in FIG. 20 to FIG. 22.

FIG. 20 is a diagram schematically showing a conventional base station 9000 and a conventional mobile station 9001 communicating with each using the DSRC system. In FIG. 20, a
15 frequency f_c denotes the center frequency for the uplink, and the value thereof is one of 5815 [MHz], 5820 [MHz], 5825 [MHz], 5830 [MHz], 5835 [MHz], 5840 [MHz] and 5845 [MHz]. Moreover, in FIG. 20, a frequency f_d denotes the difference between the center frequency of the signal used for the uplink and that of the signal
20 used for the downlink, and the value thereof is 40.000 [MHz]. As shown in FIG. 20, a signal is uplinked from the mobile station 9001 to the base station 9000 with the center frequency f_c . A signal is downlinked from the base station 9000 to the mobile station 9001 with a center frequency $f_c - f_d$. In the DSRC system, it is
25 specified that the channel bandwidth is 5 [MHz].

FIG. 21 is a diagram showing a general configuration of a conventional base-station wireless communications device employing the LOW-IF architecture. FIG. 22 is a diagram showing a general configuration of a conventional mobile-station wireless communications device employing the single-conversion architecture. For the purpose of simplifying the problem, the following description will only discuss the signal-receiving operation at the mobile-station wireless communications device and the base-station wireless communications device.

First, referring to FIG. 20 and FIG. 21, the signal-receiving operation at the base-station wireless communications device will be described. In FIG. 21, the base-station wireless communications device includes an antenna 9200, a band-pass filter 9216, a transmission/reception selector switch 9211, an amplifier 9201, a first mixer 9202, a second mixer 9203, a first local oscillator 9206, a first low-pass filter 9204, a second low-pass filter 9205, a first sampler 9207, a second sampler 9208, a sampling signal generator 9209, a demodulation digital circuit 9210, a ~~transmission high-frequency circuit 9212, a third mixer 9213, a~~ second local oscillator 9214 and a transmitter circuit 9215.

In the base-station wireless communications device, the signal-receiving operation is performed by using the antenna 9200, the band-pass filter 9216, the transmission/reception selector switch 9211, the amplifier 9201, the first mixer 9202, the second mixer 9203, the first local oscillator 9206, the first low-pass

filter 9204, the second low-pass filter 9205, the first sampler 9207, the second sampler 9208, the sampling signal generator 9209 and the demodulation digital circuit 9210.

In the signal-receiving operation, the transmission/reception selector switch 9211 is switched so that the antenna 9200 and the amplifier 9201 are connected to each other. A modulated high-frequency signal $R(t)$ from the mobile station 9001 received by the antenna 9200 whose center frequency is f_c is inputted to the amplifier 9201. The amplifier 9201 amplifies the modulated high-frequency signal $R(t)$ to an appropriate level, and inputs the amplified signal to the first mixer 9202 and the second mixer 9203. The first local oscillator 9206 outputs a sine wave whose center frequency is $f_c - f_a$. As described in Non-Patent Document 1, it is preferred that f_a is a frequency that is about a few times as great as the channel bandwidth of the modulated high-frequency signal $R(t)$.

The first mixer 9202 multiplies the sine wave outputted from the first local oscillator 9206 whose center frequency is $f_c - f_a$ with the modulated high-frequency signal $R(t)$ to output a modulated low-to-intermediate-frequency signal in-phase component $R_{XI}(t)$ whose center frequency is f_a .

The second mixer 9203 multiplies a signal outputted from the first local oscillator 9206 whose center frequency is $f_c - f_a$ and whose phase is shifted from that of the sine wave by $\pi/2$ with the modulated high-frequency signal $R(t)$ to output a modulated

low-to-intermediate-frequency signal quadrature component $RXQ(t)$ whose center frequency is f_a .

The first sampler 9207 samples the modulated low-to-intermediate-frequency signal in-phase component $RXI(t)$ in synchronism with a signal outputted from the sampling signal generator 9209 whose frequency is f_{s1} to output an in-phase component sampled signal $I(mT_{s1})$.

The second sampler 9208 samples the modulated low-to-intermediate-frequency signal quadrature component $RXQ(t)$ in synchronism with a signal outputted from the sampling signal generator 9209 whose frequency is f_{s1} to output a quadrature component sampled signal $Q(mT_{s1})$.

Herein, m is an integer, and T_{s1} is the inverse of the sampling signal frequency f_{s1} , i.e., $T_{s1} = 1/f_{s1}$. In order to facilitate the signal processing operation at the demodulation digital circuit 9210, f_{s1} is in many cases set to a value that is equal to f_a multiplied by 2^N (N is a natural number: $N=1, 2, 3, \dots$).

The demodulation digital circuit 9210 receives the in-phase component sampled signal $I(mT_{s1})$ and the quadrature component sampled signal $Q(mT_{s1})$ as input signals, and demodulates the signals to output received data after removing the image disturbing signal, as described in Non-Patent Document 1.

Next, referring to FIG. 21 and FIG. 22, the signal-receiving operation at the mobile-station wireless communications device will be described. In FIG. 22, the mobile-station wireless

communications device includes an antenna 9100, a band-pass filter 9112, a transmission/reception selector switch 9108, an amplifier 9101, a first mixer 9102, a local oscillator 9103, a low-pass filter 9104, a sampler 9105, a sampling signal generator 9106, a
5 demodulation digital circuit 9107, a transmission high-frequency circuit 9109, a second mixer 9110 and a transmitter circuit 9111.

In the mobile-station wireless communications device, the signal-receiving operation is performed by using the antenna 9100, the band-pass filter 9112, the transmission/reception selector
10 switch 9108, the amplifier 9101, the first mixer 9102, the local oscillator 9103, the low-pass filter 9104, the sampler 9105, the sampling signal generator 9106 and the demodulation digital circuit 9107.

In the signal-receiving operation, the
15 transmission/reception selector switch 9108 is switched so that the antenna 9100 and the amplifier 9101 are connected to each other. A modulated high-frequency signal $RL(t)$ from the base station 9000 received by the antenna 9100 whose center frequency is $f_c - f_d$ is
~~first passed through the band-pass filter 9112 to remove signals~~
20 of frequency bands that are used neither in the base station nor in the mobile station, and is then inputted to the amplifier 9101. The amplifier 9101 amplifies the modulated high-frequency signal $RL(t)$ to an appropriate level, and inputs the amplified signal to the first mixer 9102. The first local oscillator 9103 outputs
25 a sine wave whose center frequency is f_c .

The first mixer 9102 multiplies the sine wave outputted from the local oscillator 9103 whose center frequency is f_c with the modulated high-frequency signal $R_L(t)$ to output a modulated low-to-intermediate-frequency signal $L(t)$ whose center frequency is f_d to the low-pass filter 9104.

In the frequency conversion at the first mixer 9102, a signal whose center frequency is $f_c + f_d$ is an image disturbing signal. However, since the image response is not specified in the technical requirements for wireless equipment used in the mobile station in the DSRC system standard, a lower-order, inexpensive low-pass filter can be used as the filter following the first mixer 9102. If the image disturbing signal were to be a problem, signal components of only the necessary bands can be extracted by using a complex filter.

The sampler 9105 samples the modulated low-to-intermediate-frequency signal $L(t)$ outputted from the low-pass filter 9104 whose center frequency is f_d in synchronism with a signal outputted from the sampling signal generator 9106 whose frequency is f_{s2} to output a sampled signal $L_s(mT_{s2})$. Herein, m is an integer, and T_{s2} is a value represented by the inverse ($1/f_{s2}$) of the sampling signal frequency f_{s2} . In order to facilitate the signal processing operation at the demodulation digital circuit 9107, f_{s2} is in many cases set to a value that is equal to f_d multiplied by 2^N (N is a natural number: $N=1, 2, 3, \dots$).

The demodulation digital circuit 9107 receives the sampled signal $L_s(mT_s2)$ as an input signal, and demodulates the signal to output received data.

Other background art publications related to the present invention include Non-Patent Document 2 (Mikko Valkama, et al., "Advanced Methods for I/Q Imbalance Compensation in Communication Receivers" IEEE TRANSACTIONS ON SIGNAL PROCESSING, Vol. 49, No. 10, pp. 2335-2344, October 2001) and Non-Patent Document 3 (Kiyomichi Araki ed., "Software Musen No Kiso To Oyo (Basics and Applications of Software Radio)", SIPEC Corporation Knowledge Service Department, p. 123, October 2002).

As described above, in the base station employing the LOW-IF architecture, the center frequency f_a of the signals $R_{XI}(t)$ and $R_{XQ}(t)$ inputted to the first and second samplers 9207 and 9208 is about a few times as great as the signal bandwidth of the modulated high-frequency signal $R(t)$. In the mobile station employing the single-conversion architecture, the center frequency f_d of the signal $L(t)$ inputted to the sampler 9105 is equal to the difference ~~(40.000 [MHz]) between the uplink frequency and the downlink~~ frequency as specified in the DSRC system standard.

Therefore, the center frequency of the signal inputted to the sampler in the mobile station is substantially different from that of the signal inputted to the sampler in the base station, whereby the frequency of the sampling signal used in the sampler 9105 in the mobile station is different from that of the sampling

signal used in the first and second samplers 9207 and 9208 in the base station.

Therefore, with the conventional system, the sampling frequency for the base station and that for the mobile station need to be set to different values even though their demodulation digital circuits are substantially the same in function. Thus, it is necessary to provide two different demodulation digital circuits for the base station and for the mobile station. Although it is desirable to realize a common demodulation digital circuit for the base station and for the mobile station in order to provide an inexpensive transceiver, it is difficult to realize such a common demodulation digital circuit for reasons stated above.

DISCLOSURE OF THE INVENTION

Therefore, an object of the present invention is to realize a wireless communications system in which a common sampling frequency is used by the base station and the mobile station, thereby providing wireless digital receivers for the base station and for the mobile station at a low cost and reducing the overall cost of the wireless communications system.

The present invention has the following features to attain the object mentioned above.

A first aspect of the present invention is directed to a wireless communications system for transmitting/receiving a first wireless signal from a first wireless communications device and

a second wireless signal from a second wireless communications device, the first and second wireless signals having different frequency bands from each other, wherein: the first wireless communications device includes: a first frequency converter for downconverting the second wireless signal transmitted from the second wireless communications device to a first low-frequency signal; a first sampler for oversampling the first low-frequency signal downconverted by the first frequency converter; and a first demodulation digital circuit for demodulating the signal oversampled by the first sampler; the signal demodulated by the first demodulation digital circuit has a center frequency of f_i [Hz]; the second wireless communications device includes: a second frequency converter for downconverting the first wireless signal transmitted from the first wireless communications device to a second low-frequency signal whose center frequency f_d [Hz] is equal to a difference between a center frequency of the first wireless signal and that of the second wireless signal; a second sampler for undersampling the second low-frequency signal downconverted by the second frequency converter; and a second demodulation digital circuit for demodulating the signal undersampled by the second sampler; a sampling frequency used in the first sampler and that used in the second sampler are the same sampling frequency f_s [Hz]; the sampling frequency f_s [Hz] is set to a value that is an even-number multiple of a wireless symbol transmission rate such that oversampling is done in the first sampler and

undersampling is done in the second sampler; and the center frequency f_i [Hz] is $1/2$ to 1 times a frequency corresponding to a bandwidth of the first and second wireless signals and is $1/2^N$ (N is a natural number) times the sampling frequency f_s [Hz].

5 In a preferred embodiment, where the bandwidth of the first and second wireless signals is $2 \times B_{ch}$ [Hz] and the wireless symbol transmission rate is f_{sym} [Hz], the sampling frequency f_s [Hz] and the center frequency f_i [Hz] are expressed as shown in the following expressions:

$$f_i = \frac{2k f_{sym}}{2^N}$$

$$f_s = 2^N f_i$$

where k is an integer satisfying

$$\frac{f_d + B_{ch}}{(n+1)f_{sym}} \leq k \leq \frac{f_d - B_{ch}}{n f_{sym}} \quad \dots \text{Exp. 12}$$

and

$$k \leq \frac{f_d}{2f_{sym}} \quad \dots \text{Exp. 14}$$

and N is an integer satisfying

$$\log_2 \left\{ \frac{f_d + B_{ch}}{(n+1)B_{ch}} \right\} \leq N \leq \log_2 \left\{ \frac{2(f_d - B_{ch})}{n B_{ch}} \right\} \quad \dots \text{Exp. 22}$$

where n is an integer satisfying

$$1 \leq n \leq \frac{f_d - B_{ch}}{2B_{ch}} \quad \dots \text{Exp. 7}$$

In a preferred embodiment, the first frequency converter downconverts the second wireless signal transmitted from the second wireless communications device to a first low-frequency signal whose center frequency is f_j [Hz]; and the first low-frequency
5 signal is demodulated by the first demodulation digital circuit after being corrected to a signal whose center frequency is f_i [Hz] at a position preceding or following the first sampler.

In a preferred embodiment, the center frequency f_d is 40.000 [MHz]; and the frequency f_i and the sampling frequency f_s are
10 $f_i=3.072$ [MHz] and $f_s=24.576$ [MHz], $f_i=3.072$ [MHz] and $f_s=12.288$ [MHz], $f_i=4.608$ [MHz] and $f_s=36.864$ [MHz], $f_i=4.096$ [MHz] and $f_s=32.768$ [MHz], or $f_i=3.584$ [MHz] and $f_s=28.672$ [MHz].

In a preferred embodiment, the first demodulation digital circuit includes: a first quadrature demodulator for
15 quadrature-demodulating the signal oversampled by the first sampler; a first low-pass filter for low-pass-filtering the signal quadrature-demodulated by the first quadrature demodulator; and a first received data reproducing section for reproducing received
~~data from the signal low-pass-filtered by the first low-pass~~
20 filter; the second demodulation digital circuit includes: a second quadrature demodulator for quadrature-demodulating the signal undersampled by the second sampler; a second low-pass filter for low-pass-filtering the signal quadrature-demodulated by the
25 reproducing section for reproducing received data from the signal

low-pass-filtered by the second low-pass filter; the first quadrature demodulator converts the signal oversampled by the first sampler to a signal including a component whose center frequency is zero; and the second quadrature demodulator converts the signal undersampled by the second sampler to a signal including a component whose center frequency is zero.

In a preferred embodiment, the first demodulation digital circuit includes: a first complex filter for filtering, by using a digital filter, either one of a positive frequency component and a negative frequency component of the signal oversampled by the first sampler whose center frequency is closer to zero; and a first received data reproducing section for reproducing received data from the signal filtered by the first complex filter; and the second demodulation digital circuit includes: a second complex filter for filtering, by using a digital filter, either one of a positive frequency component and a negative frequency component of the signal undersampled by the second sampler whose center frequency is closer to zero; and a second received data reproducing section for reproducing received data from the signal filtered by the second complex filter.

In a preferred embodiment, the first demodulation digital circuit includes: a first quadrature demodulator for quadrature-demodulating the signal oversampled by the first sampler; a first low-pass filter for low-pass-filtering the signal outputted from the first quadrature demodulator; and a first

received data reproducing section for reproducing received data from the signal low-pass-filtered by the first low-pass filter; the second demodulation digital circuit includes: a second quadrature demodulator for quadrature-demodulating the signal undersampled by the second sampler; a second low-pass filter for low-pass-filtering the signal quadrature-demodulated by the second quadrature demodulator; and a second received data reproducing section for reproducing received data from the signal low-pass-filtered by the second low-pass filter; the first quadrature demodulator converts the signal oversampled by the first sampler to a signal including a component whose center frequency is zero; and the second quadrature demodulator converts the signal undersampled by the second sampler to a signal including a component whose center frequency is zero.

15 In a preferred embodiment, the frequency f_j [Hz] is 3.000 [MHz].

A second aspect of the present invention is directed to a wireless digital receiver in a wireless communications system for transmitting/receiving a first wireless signal from a first wireless communications device and a second wireless signal from a second wireless communications device, the first and second wireless signals having different frequency bands from each other, the wireless digital receiver receiving the second wireless signal in the first wireless communications device and digitally demodulating the second wireless signal, the wireless digital

receiver including: a frequency converter for downconverting the second wireless signal transmitted from the second wireless communications device to a low-frequency signal whose center frequency is f_i [Hz]; a sampler for oversampling the low-frequency signal downconverted by the frequency converter; and a demodulation digital circuit for demodulating the signal oversampled by the sampler, wherein: a sampling frequency used in the sampler and that used in the second wireless communications device are the same sampling frequency f_s [Hz]; the sampling frequency f_s [Hz] is set to a value that is an even-number multiple of a wireless symbol transmission rate such that oversampling is done in the sampler and undersampling is done in a sampler of the second wireless communications device; and the center frequency f_i [Hz] of the low-frequency signal is $1/2$ to 1 times a frequency corresponding to a bandwidth of the first and second wireless signals and is $1/2^N$ (N is a natural number) times the sampling frequency f_s [Hz].

In a preferred embodiment, where the bandwidth of the first and second wireless signals is $2 \times B_{ch}$ [Hz] and the wireless symbol transmission rate is f_{sym} [Hz], the sampling frequency f_s [Hz] and the center frequency f_i [Hz] of the low-frequency signal are expressed as shown in the following expressions:

$$f_i = \frac{2k f_{sym}}{2^N}$$

$$f_s = 2^N f_i$$

where k is an integer satisfying

$$\frac{f_d + B_{ch}}{(n+1)f_{sym}} \leq k \leq \frac{f_d - B_{ch}}{n f_{sym}} \quad \dots \text{Exp. 12}$$

and

$$k \leq \frac{f_d}{2f_{sym}} \quad \dots \text{Exp. 14}$$

5 and N is an integer satisfying

$$\log_2 \left\{ \frac{f_d + B_{ch}}{(n+1)B_{ch}} \right\} \leq N \leq \log_2 \left\{ \frac{2(f_d - B_{ch})}{n B_{ch}} \right\} \quad \dots \text{Exp. 22}$$

where n is an integer satisfying

$$1 \leq n \leq \frac{f_d - B_{ch}}{2B_{ch}} \quad \dots \text{Exp. 7}$$

In a preferred embodiment, the center frequency f_i and the
 10 sampling frequency f_s are $f_i=3.072$ [MHz] and $f_s=24.576$ [MHz],
 $f_i=3.072$ [MHz] and $f_s=12.288$ [MHz], $f_i=4.608$ [MHz] and $f_s=36.864$
 [MHz], $f_i=4.096$ [MHz] and $f_s=32.768$ [MHz], or $f_i=3.584$ [MHz] and
 $f_s=28.672$ [MHz].

In a preferred embodiment, the demodulation digital circuit
 15 includes: a quadrature demodulator for quadrature-demodulating
 the signal oversampled by the sampler; a low-pass filter for
 low-pass-filtering the signal quadrature-demodulated by the
 quadrature demodulator; and a received data reproducing section
 for reproducing received data from the signal low-pass-filtered
 20 by the low-pass filter; and the quadrature demodulator converts
 the signal oversampled by the sampler to a signal including a

component whose center frequency is zero.

In a preferred embodiment, the demodulation digital circuit includes: a complex filter for filtering, by using a digital filter, either one of a positive frequency component and a negative
5 frequency component of the signal oversampled by the sampler whose center frequency is closer to zero; and a received data reproducing section for reproducing received data from the signal filtered by the complex filter.

A third aspect of the present invention is directed to a
10 wireless digital receiver in a wireless communications system for transmitting/receiving a first wireless signal from a first wireless communications device and a second wireless signal from a second wireless communications device, the first and second wireless signals having different frequency bands from each other,
15 the wireless digital receiver receiving the first wireless signal in the second wireless communications device and digitally demodulating the first wireless signal, the wireless digital receiver including: a frequency converter for downconverting the
first wireless signal transmitted from the first wireless
20 communications device to a low-frequency signal whose center frequency f_d [Hz] is equal to a difference between a center frequency of the first wireless signal and that of the second wireless signal; a sampler for undersampling the low-frequency signal downconverted by the frequency converter; and a demodulation digital circuit
25 for demodulating the signal undersampled by the sampler, wherein:

a sampling frequency used in the sampler and that used in the first wireless communications device are the same sampling frequency f_s [Hz]; and the sampling frequency f_s [Hz] is set to a value that is an even-number multiple of a wireless symbol transmission rate
 5 such that undersampling is done in the sampler and oversampling is done in a sampler of the first wireless communications device.

In a preferred embodiment, where the bandwidth of the first and second wireless signals is $2 \times B_{ch}$ [Hz] and the wireless symbol transmission rate is f_{sym} [Hz], the sampling frequency f_s [Hz]
 10 is expressed as shown in the following expression:

$$f_s = 2k f_{sym}$$

where k is an integer satisfying

$$\frac{f_d + B_{ch}}{(n+1)f_{sym}} \leq k \leq \frac{f_d - B_{ch}}{n f_{sym}} \quad \dots \text{Exp. 12}$$

and

$$k \leq \frac{f_d}{2f_{sym}} \quad \dots \text{Exp. 14}$$

15

where n is an integer satisfying

$$1 \leq n \leq \frac{f_d - B_{ch}}{2B_{ch}} \quad \dots \text{Exp. 7}$$

In a preferred embodiment, the center frequency f_d is 40.000 [MHz]; and the sampling frequency f_s is 24.576 [MHz], 12.288 [MHz],
 20 $f_s = 36.864$ [MHz], $f_s = 32.768$ [MHz] or $f_s = 28.672$ [MHz].

In a preferred embodiment, the demodulation digital circuit includes: a quadrature demodulator for quadrature-demodulating

the signal undersampled by the sampler; and a low-pass filter for low-pass-filtering the signal quadrature-demodulated by the quadrature demodulator; and a received data reproducing section for reproducing received data from the signal low-pass-filtered
5 by the low-pass filter; and the quadrature demodulator converts the signal undersampled by the sampler to a signal including a component whose center frequency is zero.

In a preferred embodiment, the demodulation digital circuit includes: a complex filter for filtering, by using a digital filter,
10 either one of a positive frequency component and a negative frequency component of the signal undersampled by the sampler whose center frequency is closer to zero; and a received data reproducing section for reproducing received data from the signal filtered by the complex filter.

15 A fourth aspect of the present invention is directed to a wireless digital receiver in a wireless communications system for transmitting/receiving a first wireless signal from a first wireless communications device and a second wireless signal from
a second wireless communications device, the first and second
20 wireless signals having different frequency bands from each other, the wireless digital receiver receiving the second wireless signal in the first wireless communications device and digitally demodulating the second wireless signal, the wireless digital receiver including: a frequency converter for downconverting the
25 second wireless signal transmitted from the second wireless

communications device to a low-frequency signal whose center frequency is f_j [Hz]; a sampler for oversampling the low-frequency signal downconverted by the frequency converter; and a demodulation digital circuit for demodulating the signal oversampled by the
 5 sampler after correcting a center frequency thereof to f_i [Hz], wherein: a sampling frequency used in the sampler and that used in the second wireless communications device are the same sampling frequency f_s [Hz]; the sampling frequency f_s [Hz] is set to a value that is an even-number multiple of a wireless symbol transmission
 10 rate such that oversampling is done in the sampler and undersampling is done in a sampler of the second wireless communications device; and the center frequency f_i [Hz] is $1/2$ to 1 times a frequency corresponding to a bandwidth of the first and second wireless signals and is $1/2^N$ (N is a natural number) times the sampling
 15 frequency f_s [Hz].

In a preferred embodiment, where the bandwidth of the first and second wireless signals is $2 \times B_{ch}$ [Hz] and the wireless symbol transmission rate is f_{sym} [Hz], the sampling frequency f_s [Hz] and the frequency f_i [Hz] are expressed as shown in the following
 20 expressions:

$$f_i = \frac{2k f_{sym}}{2^N}$$

$$f_s = 2^N f_i$$

where k is an integer satisfying

$$\frac{fd+Bch}{(n+1)fsym} \leq k \leq \frac{fd-Bch}{nfsym} \quad \dots \text{Exp. 12}$$

and

$$k \leq \frac{fd}{2fsym} \quad \dots \text{Exp. 14}$$

and N is an integer satisfying

$$\log_2 \left\{ \frac{fd+Bch}{(n+1)Bch} \right\} \leq N \leq \log_2 \left\{ \frac{2(fd-Bch)}{nBch} \right\} \quad \dots \text{Exp. 22}$$

where n is an integer satisfying

$$1 \leq n \leq \frac{fd-Bch}{2Bch} \quad \dots \text{Exp. 7}$$

In a preferred embodiment, the demodulation digital circuit includes: a quadrature demodulator for quadrature-demodulating the signal oversampled by the sampler; an automatic frequency controller for correcting the signal quadrature-demodulated by the quadrature demodulator to a signal having a component whose frequency is f_i [Hz]; a low-pass filter for low-pass-filtering the signal frequency-corrected by the automatic frequency controller; and a received data reproducing section for reproducing received data from the signal low-pass-filtered by the low-pass filter.

In a preferred embodiment, the frequency f_j [Hz] is 3.000 [MHz].

Each functional block of the present invention is preferably implemented in the form of an integrated circuit. The integrated

circuits implementing these functional blocks may be individually formed into a separate chip, or some or all of them may be formed together into a single chip. In the present specification, "an integrated circuit" refers not only to an integrated circuit
5 provided in the form of a single chip, but also to a group of integrated circuits that are formed together into a single chip.

The effects of the present invention will now be described. In a wireless communications system of the present invention and a wireless digital receiver for use therein, the sampling frequency
10 used in the first wireless communications device (base station) is the same as that used in the second wireless communications device (mobile station). Therefore, the same demodulation digital circuit for performing a digital demodulation operation can be used for the first and second wireless communications devices.
15 Therefore, with the present invention, it is not necessary to provide a separate demodulation digital circuit for the first and second wireless communications devices (the base station and the mobile station), whereby it is possible to provide an inexpensive
wireless digital receiver, thereby reducing the overall cost of
20 the wireless communications system.

Moreover, with the provision of the automatic frequency controller in the demodulation digital circuit, the local oscillator is allowed a certain degree of freedom, which also contributes to reducing the cost.

25 These and other objects, features, aspects and advantages

of the present invention will become more apparent from the following detailed description of the present invention when taken in conjunction with the accompanying drawings.

5

BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a block diagram showing a functional configuration of a wireless communications system 1 according to a first embodiment of the present invention;

10

FIG. 2 is a block diagram showing a functional configuration of a first wireless digital receiver 21;

FIG. 3 is a diagram showing pass-band characteristics of a low-pass filter 103;

FIG. 4 is a block diagram showing a functional configuration of a second wireless digital receiver 31;

15

FIG. 5 is a diagram showing a configuration of a quadrature modulator;

FIG. 6A and FIG. 6B are diagrams showing the spectrum of a transmitted signal, and the result of a multiplication of the transmitted signal with a sine wave;

20

FIG. 7A and FIG. 7B are diagrams showing the spectrum of a transmitted signal, and the result of a multiplication of the transmitted signal with a sine wave;

FIG. 8 is a diagram showing the spectrum of a sampled signal $S_1(mT_s)$ outputted from a sampler 101;

25

FIG. 9 is a diagram showing a configuration of a quadrature

demodulator;

FIG. 10 is a diagram showing the spectrum of a sampled signal $S2(mTs)$ obtained by sampling a modulated low-frequency signal $L2(t)$ whose center frequency is $f_d=40.000$ [MHz] with the sampling frequency $f_s=24.576$ [MHz];

FIG. 11 is a diagram showing the spectrum of the sampled signal $S1(mTs)$ outputted from the sampler 101 where the center frequency of a modulated low-frequency signal $L1(t)$ is $f_i=3.072$ [MHz] and the sampling frequency f_s is set to 12.288 [MHz];

10 FIG. 12 is a diagram showing the spectrum of the sampled signal $S2(mTs)$ obtained by sampling the modulated low-frequency signal $L2(t)$ whose center frequency is $f_d=40.000$ [MHz] with the sampling frequency $f_s=12.288$ [MHz];

FIG. 13 is a block diagram showing a functional configuration of the first wireless digital receiver 21 according to a second embodiment of the present invention;

FIG. 14 is a diagram showing exemplary pass-band characteristics of a complex filter 602;

FIG. 15A, FIG. 15B, FIG. 15C, FIG. 15D, FIG. 15E and FIG. 15F are diagrams used for illustrating the pass-band characteristics of the complex filter 602;

FIG. 16A and FIG. 16B are block diagrams each showing a functional configuration of the first wireless digital receiver 21 according to a third embodiment of the present invention;

25 FIG. 17 is a diagram showing the spectrum of an in-phase

component sampled signal I (mTs) and a quadrature component sampled signal Q (mTs) outputted from a quadrature demodulator 802;

FIG. 18 is a diagram showing a configuration of a base-station wireless communications device 12 according to a fourth embodiment
5 of the present invention;

FIG. 19 is a diagram showing a configuration of a mobile-station wireless communications device 11 according to the fourth embodiment of the present invention;

FIG. 20 is a diagram schematically showing a conventional
10 base station 9000 and a conventional mobile station 9001 communicating with each other using the DSRC system;

FIG. 21 is a diagram showing a general configuration of a conventional base-station wireless communications device employing the LOW-IF architecture; and

15 FIG. 22 is a diagram showing a general configuration of a conventional mobile-station wireless communications device employing the single-conversion architecture.

BEST MODE FOR CARRYING OUT THE INVENTION

20 (First Embodiment)

FIG. 1 is a block diagram showing a functional configuration of a wireless communications system 1 according to a first embodiment of the present invention. In FIG. 1, the wireless communications system 1 includes a base station 2 being a first
25 wireless communications device, and a mobile station 3 being a

second wireless communications device. The base station 2 includes a first wireless digital receiver 21 and a first wireless transmitter 22. The mobile station 3 includes a second wireless digital receiver 31 and a second wireless transmitter 32. FIG. 1 only shows one base station 2 and one mobile station 3 for the sake of simplicity. In practice, however, there are a plurality of base stations 2 and a plurality of mobile stations 3 communicating with one another using different channels.

The wireless communications system 1 employs the DSRC system standard. Therefore, one of 5815 [MHz], 5820 [MHz], 5825 [MHz], 5830 [MHz], 5835 [MHz], 5840 [MHz] and 5845 [MHz] is used for the uplink from the mobile station 3 to the base station 2. Herein, the center frequency of a signal used for the uplink is denoted as f_c [Hz] (hereinafter referred to simply as " f_c ").

For the downlink from the base station 2 to the mobile station 3, one of 5775 [MHz], 5780 [MHz], 5785 [MHz], 5790 [MHz], 5795 [MHz], 5800 [MHz] and 5805 [MHz] is used according to the frequency f_c for the uplink. The frequency f_d [Hz] (hereinafter referred to simply as " f_d "), which is equal to the difference between the frequency used for the downlink and that used for the uplink, is always 40.000 [MHz]. The center frequency of a signal used for the downlink is $f_c - f_d$ [Hz] (hereinafter referred to simply as " $f_c - f_d$ ").

The DSRC system standard specifications include a section on the technical requirements for wireless equipment, which

specifies various requirements for the base station and the mobile station.

The DSRC system standard specifies that the bandwidth of a signal in each channel (hereinafter referred to as the "channel bandwidth") is 5 [MHz]. According to this specification, it is understood that where the channel bandwidth is denoted as $2 \times B_{ch}$, $B_{ch} = 2.5$ [MHz].

In the DSRC system, it is specified that the modulation scheme employed needs to be either the ASK (Amplitude Shift Keying) scheme in which the wireless symbol frequency f_{sym} [Hz] (hereinafter referred to simply as " f_{sym} ") is 1.024 [MHz] or the $\pi/4$ shift QPSK (Quadrature Phase Shift Keying) in which the wireless symbol frequency f_{sym} is 2.048 [MHz]. The DSRC system of the present embodiment employs the $\pi/4$ shift QPSK scheme with the wireless symbol frequency $f_{sym} = 2.048$ [MHz]. In the case of the ASK scheme, since the Manchester coding is used, $f_{sym} = 1.024$ [MHz] converted to a baud rate is 2.048 [MHz], which is equal to that of the $\pi/4$ shift QPSK scheme. Therefore, the following description in principle applies to the ASK scheme.

The first wireless transmitter 22 of the base station 2 outputs a signal (first wireless signal) whose center frequency is $f_c - f_d$. In response, the second wireless digital receiver 31 of the mobile station 3 receives the signal (first wireless signal) whose center frequency is $f_c - f_d$. The second wireless digital receiver 31 downconverts the received signal (first wireless

signal) whose center frequency is $f_c - f_d$ to a signal whose center frequency is $f_d = 40.000$ [MHz]. The second wireless digital receiver 31 undersamples the signal whose center frequency is f_d in synchronism with the sampling signal whose sampling frequency is $f_s = 24.576$ [MHz]. The second wireless digital receiver 31 demodulates the undersampled signal by using a digital circuit to obtain received data.

The second wireless transmitter 32 of the mobile station 3 outputs a signal (second wireless signal) whose center frequency is f_c . In response, the first wireless digital receiver 21 of the base station 2 receives the signal (second wireless signal) whose center frequency is f_c . The first wireless digital receiver 21 downconverts the received signal (second wireless signal) whose center frequency is f_c to a signal whose center frequency is $f_i = 3.072$ [MHz]. The first wireless digital receiver 21 oversamples the signal whose center frequency is f_i in synchronism with the sampling signal whose sampling frequency is $f_s = 24.576$ [MHz]. The first wireless digital receiver 21 demodulates the oversampled signal by using a digital circuit to obtain received data.

FIG. 2 is a block diagram showing a functional configuration of the first wireless digital receiver 21. In FIG. 2, the first wireless digital receiver 21 includes a frequency converter 100, a sampler 101, a quadrature demodulator 102, a low-pass filter 103, a sampling signal generator 104 and a received data reproducing section 105. The quadrature demodulator 102, the low-pass filter

103 and the received data reproducing section 105 will be hereinafter referred to collectively as a "first demodulation digital circuit". Assume that a modulated high-frequency signal $R1(t)$ whose center frequency is f_c is inputted to the first wireless
5 digital receiver 21.

The frequency converter 100 downconverts the modulated high-frequency signal $R1(t)$ and outputs the modulated low-frequency signal $L1(t)$ whose center frequency is $f_i = 3.072$ [MHz]. The reason why the signal is downconverted to $f_i = 3.072$ [MHz] will
10 later be described in detail.

The sampling signal generator 104 outputs a sampling signal whose sampling frequency is $f_s = 24.576$ [MHz]. The reason why the sampling frequency is $f_s = 24.576$ [MHz] will later be described in detail.

15 The sampler 101 oversamples the modulated low-frequency signal $L1(t)$ in synchronism with the sampling signal outputted from the sampling signal generator 104 to output a sampled signal $S1(mT_s)$. Herein, m is an integer ($m = \dots, -1, 0, 1, \dots$), and T_s is the sampling period, i.e., $T_s = 1/f_s$.

20 The quadrature demodulator 102 performs an operation of multiplying the sampled signal $S1(mT_s)$ outputted from the sampler 101 by $\exp(-j\theta \times mT_s)$ (where j is the imaginary unit) to output two signals whose phases are different from each other by $\pi/2$ [rad], i.e., an in-phase component sampled signal $I1(mT_s)$ and a quadrature
25 component sampled signal $Q1(mT_s)$. Herein, θ is set to a value

such that a signal that is frequency-shifted so that the center frequency thereof is zero is included in the signals outputted from the quadrature demodulator 102 after the multiplication with $\exp(-j\theta \times mTs)$. The value θ will later be described in detail.

5 FIG. 3 is a diagram showing pass-band characteristics of the low-pass filter 103. A low-pass filter 203 is a digital filter whose frequency pass band is zero to $Bch/2$. With the provision of the low-pass filter 103, the in-phase component $Ib1(mTs)$ and the quadrature component $Qb1(mTs)$ of the baseband quadrature
10 demodulated signal outputted from the low-pass filter 103 will only have a frequency component that is frequency-shifted so that the center frequency thereof is zero.

The signals $Ib1(mTs)$ and $Qb1(mTs)$ outputted from the low-pass filter 103 only have a component that is frequency-shifted so that
15 the center frequency thereof is zero, whereby the received data reproducing section 105 can output the received data by means of delay detection, or the like.

FIG. 4 is a block diagram showing a functional configuration of the second wireless digital receiver 31. In FIG. 4, the second
20 wireless digital receiver 31 includes a frequency converter 200, a sampler 201, a quadrature demodulator 202, the low-pass filter 203, a sampling signal generator 204 and a received data reproducing section 205. The quadrature demodulator 202, the low-pass filter 203 and the received data reproducing section 205 will be
25 hereinafter referred to collectively as a "second demodulation

digital circuit". Assume that a modulated high-frequency signal $R2(t)$ whose center frequency is $f_c - f_d$ is inputted to the second wireless digital receiver 31.

The frequency converter 200 downconverts the modulated high-frequency signal $R2(t)$ to output the modulated low-frequency signal $L2(t)$ whose center frequency is $f_d = 40.000$ [MHz]. In the mobile station of the DSRC system, a local oscillator signal whose frequency is f_c is outputted from a local oscillator (not shown) in order to output a signal to be transmitted. The mobile station employs a single-conversion architecture using the local oscillator signal. Moreover, the frequency of the signal that the mobile station receives is $f_c - f_d$. Therefore, the frequency converter 200 downconverts the modulated high-frequency signal $R2(t)$ to $f_d = 40.000$ [MHz] by using the local oscillator signal whose frequency is f_c .

The sampling signal generator 204 outputs a sampling signal whose sampling frequency is $f_s = 24.576$ [MHz]. Thus, the sampling signal generator 204 is the same as the sampling signal generator 104 in the first wireless digital receiver 21. As described above, in the present embodiment, the sampling frequency used in the first wireless digital receiver 21 is equal to that used in the second wireless digital receiver 31. The reason why the same sampling frequency can be used will later be described. Also, the reason why the sampling frequency is $f_s = 24.576$ [MHz] will later be described in detail.

The sampler 201 undersamples the modulated low-frequency signal $L2(t)$ in synchronism with the sampling signal outputted from the sampling signal generator 204 to output a sampled signal $S2(mT_s)$. Herein, m is an integer ($m = \dots, -1, 0, 1, \dots$), and T_s is the sampling period, i.e., $T_s = 1/f_s$.

The quadrature demodulator 202 performs an operation of multiplying the sampled signal $S2(mT_s)$ outputted from the sampler 201 by $\exp(-j\eta \times mT_s)$ (where j is the imaginary unit) to output two signals whose phases are different from each other by $\pi/2$ [rad], i.e., an in-phase component sampled signal $I2(mT_s)$ and a quadrature component sampled signal $Q2(mT_s)$. Herein, η is set to a value such that a signal that is frequency-shifted so that the center frequency thereof is zero is included in the signals outputted from the quadrature demodulator 202 after the multiplication with $\exp(-j\eta \times mT_s)$. The value η will later be described in detail. As will be described later, η is a value different from θ used in the quadrature demodulator 202 of the first wireless digital receiver 21. Thus, the quadrature demodulator 102 used in the first wireless digital receiver 21 is the same as the quadrature demodulator 202 used in the second wireless digital receiver 31 except that the rotation angles θ and η used in the multiplication with \exp are different from each other. The rotation angles being different from each other will later be described in detail.

The low-pass filter 203 is a digital filter whose frequency pass band is zero to $B_{ch}/2$, as is the low-pass filter 103 in the

first wireless digital receiver 21. Thus, FIG. 3 is relied upon also for the low-pass filter 203. With the provision of the low-pass filter 203, the in-phase component $I_{b2}(mTs)$ and the quadrature component $Q_{b2}(mTs)$ of the baseband quadrature demodulated signal outputted from the low-pass filter 203 will only have a frequency component that is frequency-shifted so that the center frequency thereof is zero. The low-pass filter 103 used in the first wireless digital receiver 21 is the same as the low-pass filter 203 used in the second wireless digital receiver 31.

The signals $I_{b2}(mTs)$ and $Q_{b2}(mTs)$ outputted from the low-pass filter 203 only have a component that is frequency-shifted so that the center frequency thereof is zero, whereby the received data reproducing section 205 can output the received data by means of delay detection, or the like.

Since the sampling frequency for the first wireless digital receiver 21 is equal to that for the second wireless digital receiver 31, the same sampling signal generator can be used as the sampling signal generators 104 and 204. Moreover, the same sampler can be used as the samplers 101 and 201. Furthermore, the same low-pass filter can be used as the low-pass filters 103 and 203. In addition, since it is only necessary to change the rotation angle between the quadrature demodulator 102 and the quadrature demodulator 202, the same quadrature demodulator can be used as the quadrature demodulator 102 in the first wireless digital receiver 21 and the

quadrature demodulator 202 in the second wireless digital receiver 31 if two different rotation angles can be stored in a memory device and the quadrature demodulator used is capable of switching the rotation angle values from one to another.

5 Now, the reason why sampling is done properly at the samplers 101 and 201 and the received data can be properly obtained at the first and second wireless digital receivers 21 and 31 by using $f_i=3.072$ [MHz] and $f_s=24.576$ [MHz] will be described. Specifically the reason why the received data can be restored
10 completely as the sampler 101 oversamples a signal whose center frequency is $f_i=3.072$ [MHz] with the sampling frequency $f_s=24.576$ [MHz] while the received data can be restored completely as the sampler 201 undersamples a signal whose center frequency is $f_i=40.000$ [MHz] with the sampling frequency $f_s=24.576$ [MHz] will
15 be described.

A transmitted signal can generally be expressed as shown in Expression 1 below, using a complex signal.

$$\text{Re}[S(t)\exp\{j(\omega_c t + \phi)\}] \quad \dots \text{Exp. 1}$$

~~This is because the transmitted baseband signal $S(t)$ is,~~
20 in the first place, a complex signal expressed as $T_x I + j T_x Q$, which is quadrature-modulated (multiplied with Expression 2) by using a quadrature modulator as shown in FIG. 5 and then outputted as a radio wave.

$$\exp\{j(\omega_c t + \phi)\} \quad \dots \text{Exp. 2}$$

25 The receiving side downconverts the transmitted signal by

multiplying it with a sine wave. First, the transmitted signal and the sine wave can be expressed by using a complex signal as shown in Expression 3 and Expression 4, respectively.

Transmitted Signal

$$\begin{aligned} \text{Re}[S(t)\exp\{j(\omega_c t + \phi)\}] &= \frac{1}{2} \{S(t)\exp\{j(\omega_c t + \phi)\} \\ &\quad + S^*(t)\exp\{j(\omega_c t + \phi)\}^*\} \end{aligned}$$

... Exp. 3

Sine Wave

$$\begin{aligned} \cos\{(\omega_c - \omega_i)t + \phi\} &= \frac{1}{2} \{\exp[j\{(\omega_c - \omega_i)t + \phi\}] \\ &\quad + \exp[j\{(\omega_c - \omega_i)t + \phi\}]^*\} \end{aligned}$$

...Exp. 4

5

According to Expression 3, the spectrum of the transmitted signal can be expressed on a plane as shown in FIG. 6A where the horizontal axis represents the complex frequency and the vertical axis represents the spectral intensity.

10

It can be seen from FIG. 6A that the transmitted signal is a signal made up of a spectrum of $S(t)$ at a center angular frequency of $+\omega_c$ and another spectrum of $S^*(t)$ at a center angular frequency of $-\omega_c$.

Similarly, it can be seen from Expression 4 that the sine wave is a signal made up of a sine-wave signal whose center angular frequency is $+\omega_c$ and another sine-wave signal whose center angular frequency is $-\omega_c$.

Where the center angular frequency of the local oscillator used for the downconversion is $\omega_c - \omega_i$, the frequency-converted

signal obtained by multiplying the transmitted signal by the sine wave can be expressed as shown in Expression 5.

$$\begin{aligned} & \text{Re}[S(t)\exp\{j(\omega_c t + \phi)\}]\cos\{(\omega_c - \omega_i)t + \phi\} \\ &= \frac{1}{4}[S(t)\exp\{j(\omega_i t + \phi - \phi)\} + S(t)\exp\{j(\omega_i t + \phi - \phi)\}^*] \end{aligned}$$

...Exp. 5

FIG. 6B diagrammatically shows the downconversion represented by Expression 5. It can be seen that when the transmitted signal is frequency-converted so that the center angular frequency is ω_i (a frequency value as close to zero as possible while the intended wave does not contain a DC component), an adjacent channel (ch1-) falls into the band of the intended wave, which thus becomes a disturbing wave. In principle, ch1- can be removed by using, for example, an image rejection mixer (see p. 281 of Non-Patent Document 1). In practice, however, the suppression can only be done by about 30 to 40 dB at best due to a quadrature error between the in-phase component and the quadrature component of the quadrature demodulated signal, as known in the art (see Non-Patent Document 2). However, since the adjacent wave selectivity at 5 [MHz] intervals is specified in the DSRC system neither for the base station nor for the mobile station (see STD-T75, Ver. 1.2, P. 33), ch1- does not have to be suppressed completely.

FIG. 7B shows the result of gradually moving away the center angular frequency ω_i in the positive direction from its position shown in FIG. 6B. As shown in FIG. 7B, the next adjacent channel

ch2- (the next adjacent channel is defined as a 10 [MHz]-interval signal in STD-T75, Ver. 1.2, P. 33) falls into the band of the intended wave. In such a case, if the ch2-removal deteriorates even by a small degree, the margin from the standard 15 dB will
 5 decrease, and the standard will no longer be met in worst cases. Therefore, it is preferred that the center angular frequency is as close to zero as possible.

In view of the above, when employing a LOW-IF architecture for the base station of the DSRC system, it is preferred that settings
 10 are made so that the adjacent channel ch1-* falls into the intended wave band. Thus, where the center frequency of the downconverted signal is f_i , Expression 6 below should be satisfied.

$$B_{ch} \leq f_i \leq 2B_{ch} \quad \dots \text{Exp. 6}$$

where $f_i = 2\pi \omega_i$ and $2B_{ch}$ is bandwidth per channel

Since $2 \times B_{ch} = 5$ [MHz] and $f_d = 40.000$ [MHz], a comparison between
 15 f_i and f_d yields $f_i < f_d$. Therefore, when a signal whose center frequency is f_i and another signal whose center frequency is f_d are to be sampled with the same sampling frequency, the signal whose center frequency is f_i will be oversampled while the signal whose center frequency is f_d will be undersampled. A sampling
 20 frequency such that the signals will both be oversampled can be used. In such a case, however, the sampling frequency will be very high, and it will be necessary to use a sampler capable of handling high-frequency signals, thereby making it difficult to realize the circuit at a low cost.

Therefore, the signal to be undersampled is a signal whose center frequency is f_d . A necessary and sufficient condition for realizing the undersampling operation is Expression 7 and Expression 8 below (see Non-Patent Document 3, p. 123, Expressions 5 B.12 and B.16).

$$1 \leq n \leq \frac{f_d - B_{ch}}{2B_{ch}} \quad \dots \text{Exp. 7}$$

$$\frac{2(f_d + B_{ch})}{n+1} \leq f_s \leq \frac{2(f_d - B_{ch})}{n} \quad \dots \text{Exp. 8}$$

Herein, f_s represents the sampling frequency.

The signal to be oversampled is a signal whose center frequency is f_i . According to the Nyquist's theorem, a necessary and sufficient condition for realizing the oversampling operation is Expression 9 below.

$$f_s \geq 2B_{ch} \quad \dots \text{Exp. 9}$$

Moreover, the condition for easily realizing the demodulation digital circuit is generally represented by Expression 10 below.

$$f_s = 2^N f_i = 2k f_{sym} \quad \dots \text{Exp. 10}$$

Herein, N and k are integers, and f_{sym} is a frequency representing the wireless symbol transmission rate.

Expression 8 and Expression 10 yield Expression 11.

$$\frac{2(f_d + B_{ch})}{n+1} \leq 2k f_{sym} \leq \frac{2(f_d - B_{ch})}{n} \quad \dots \text{Exp. 11}$$

Expression 11 can be rearranged with respect to k, yielding Expression 12.

$$\frac{f_d + B_{ch}}{(n+1)f_{sym}} \leq k \leq \frac{f_d - B_{ch}}{nf_{sym}} \quad \dots \text{Exp. 12}$$

The conditions for k will be further discussed. As described above, the undersampling scheme is used for a signal whose center frequency is f_d . Considering this fact together with Expression 10 yields Expression 13.

$$f_d \geq f_s = 2kf_{sym} \quad \dots \text{Exp. 13}$$

This can be transformed with respect to k, yielding Expression 14.

$$k \leq \frac{f_d}{2f_{sym}} \quad \dots \text{Exp. 14}$$

Note that since Expression 15 holds true except when $n=1$ (i.e., it holds true when $n \geq 2$), Expression 14 is an expression that always holds true as long as Expression 11 holds true.

$$f_d > \frac{2(f_d - B_{ch})}{n} \quad \dots \text{Exp. 15}$$

15

Rearranging Expression 6 by using Expression 10 yields Expression 16.

$$2^N B_{ch} \leq f_s \leq 2^{N+1} B_{ch} \quad \dots \text{Exp. 16}$$

Next, the condition under which Expression 8 and Expression 16 are satisfied at the same time will be considered. First, the condition under which Expression 8 and Expression 16 are not

satisfied at the same time will be considered. The condition under which there is no solution that satisfies Expression 8 and Expression 16 at the same time is as shown in Expression 17.

$$2^{N+1}Bch < \frac{2(fd+Bch)}{n+1} \text{ or } 2^NBch > \frac{2(fd-Bch)}{n} \quad \dots \text{Exp. 17}$$

5 Rearranging Expression 17 with respect to N yields Expression 18.

$$N < \log_2 \left\{ \frac{fd+Bch}{(n+1)Bch} \right\} \text{ or } N > \log_2 \left\{ \frac{2(fd-Bch)}{nBch} \right\} \quad \dots \text{Exp. 18}$$

Now, a calculation as shown in Expression 19 below is done for a comparison between antilogarithms in Expression 18.

$$\begin{aligned} \frac{2(fd-Bch)}{nBch} - \frac{fd+Bch}{(n+1)Bch} &= \frac{(n+1)fd - (3n+1)Bch}{n(n+1)Bch} \\ &= \frac{fd - (3 - \frac{2}{n+1})Bch}{nBch} \quad \dots \text{Exp. 19} \end{aligned}$$

10

Since $1 \leq n$ based on Expression 7, $2/(n+1) \leq 1$. Thus, Expression 20 can be obtained from Expression 19.

$$\frac{fd-3Bch}{nBch} < \frac{fd - (3 - \frac{2}{n+1})Bch}{nBch} \leq \frac{fd-2Bch}{nBch} \quad \dots \text{Exp. 20}$$

15 Since $fd=40.000$ [MHz] $=16Bch$ in the DSRC system, the value of Expression 20 is greater than zero. Thus, Expression 21 holds true.

$$\frac{2(fd-Bch)}{nBch} > \frac{fd+Bch}{(n+1)Bch} \quad \dots \text{Exp. 21}$$

Therefore, the condition for satisfying Expression 8 and Expression 16 at the same time can be obtained by negating Expression 18 as shown in Expression 22.

$$\log_2\left\{\frac{f_d+B_{ch}}{(n+1)B_{ch}}\right\} \leq N \leq \log_2\left\{\frac{2(f_d-B_{ch})}{nB_{ch}}\right\} \quad \dots \text{Exp. 22}$$

5 Thus, values of f_i and f_s of the present invention are obtained by first obtaining a value of n that satisfies Expression 7. Then, the value of k that satisfies Expression 12 and Expression 14 is obtained. Then, for the obtained value of n , the value of N that satisfies Expression 22 is obtained. Then, the obtained values
10 of N and k are substituted into Expression 10 to obtain f_i . Then, the obtained values of N and f_i are substituted into Expression 10 to obtain f_s .

Now, values of f_i and f_s will be obtained with an actual DSRC system. In the DSRC system, it is presumed that $B_{ch}=2.5$ [MHz],
15 $f_d=40.000$ [MHz] and $f_{sym}=2.048$ [MHz].

First, an integer that satisfies Expression 7 is derived. In the illustrated example, $n=1, 2, \dots, 7$ satisfies Expression
7.

Then, one of the integer values of n is selected, and an
20 integer k that satisfies Expression 12 and Expression 14 is derived. With a certain integer n ($1 \leq n \leq 7$), there may possibly be no integer k that satisfies Expression 12 and Expression 14. Specifically, when $n=1, 4, 5$ and 7 , there is no integer k that satisfies Expression 12 and Expression 14. When $n=2$, $k=7, 8$ or 9 . When $n=3$, $k=6$. When

$n=6, k=3$.

Then, one of the integer values n ($1 \leq n \leq 7$) satisfying Expression 7 is selected, and an integer N that satisfies Expression 22 is derived. When $n=1$, $N=4$. When $n=2$ or 3 , $N=3$. When $n=4, 5, 6$ or 7 , $N=2$.

Finally, based on Expression 10, the value f_i for the integers k and N is obtained, based on which the value f_s is obtained.

Table 1 below shows possible combinations of n , k and N , and the values f_i and f_s therefor.

10 TABLE 1

n	N	k	f_i [MHz]	f_s [MHz]
1	4	—	—	—
2	3	9	4.608	36.864
2	3	7	3.584	28.672
2	3	8	4.096	32.768
3	3	6	3.072	24.576
4	2	—	—	—
5	2	—	—	—
6	2	3	3.072	12.288
7	2	—	—	—

In Table 1, "—" means that there is no value that satisfies the conditions described above.

As can be seen from Table 1, the minimum value for f_i is 3.072 [MHz], and as can be seen from the description above with reference to FIG. 6A, FIG. 6B, FIG. 7A and FIG. 7B, the falling of the next adjacent channel ch_2 into the intended wave band is

least significant when f_i is 3.072 [MHz]. Thus, the following description will be limited to a case where f_i is 3.072 [MHz]. When f_i is 3.072 [MHz], f_s is 24.576 [MHz] or 12.288 [MHz] based on Table 1. In the present embodiment, 24.576 [MHz] is used as
5 f_s .

The description above shows that the received data can be restored completely as the sampler 101 oversamples a signal whose center frequency is $f_i=3.072$ [MHz] with the sampling frequency $f_s=24.576$ [MHz] while the received data can be restored completely
10 as the sampler 201 undersamples a signal whose center frequency is $f_i=40.000$ [MHz] with the sampling frequency $f_s=24.576$ [MHz]. While $f_i=3.072$ [MHz] and $f_s=24.576$ [MHz] in the illustrated example, a combination of $f_i=3.072$ [MHz] and $f_s=12.288$ [MHz] may also be used as can be seen from Table 1. Moreover, as can be seen from
15 Table 1, other possible combinations include: $f_i=4.608$ [MHz] and $f_s=36.864$ [MHz]; $f_i=4.096$ [MHz] and $f_s=36.768$ [MHz]; and $f_i=3.584$ [MHz] and $f_s=28.672$ [MHz]. Note that $f_d=40.000$ [MHz] in any case.

FIG. 8 is a diagram showing the spectrum of the sampled signal S_1 (mTs) outputted from the sampler 101. In FIG. 8, the horizontal
20 axis represents the complex frequency and the vertical axis represents the power spectral intensity.

In FIG. 8, $2B_{ch}$ represents the channel bandwidth, and $2B_{ch}=5$ [MHz] in the DSRC system. In FIG. 8, a spectrum 300 represents the spectrum of the modulated low-frequency signal $L_1(t)$. The
25 other spectra are folding spectra occurring as a result of sampling

the modulated low-frequency signal $L1(t)$ with a sampling period of T_s . The figure shows, as folding spectra, a signal whose center frequency is $f_s \pm f_i$ and another signal whose center frequency is $-f_s \pm f_i$.

5 The quadrature demodulator 102 receives the sampled signal $S1(mT_s)$ outputted from the sampler 101, and outputs two signals whose phases are different from each other by $\pi/2$ [rad], i.e., the in-phase component sampled signal $I1(mT_s)$ and the quadrature component sampled signal $Q1(mT_s)$. Specifically, the quadrature
10 demodulator 102 performs an operation $S1(mT_s) \times \exp(-j\theta \times mT_s)$ by using θ [rad] expressed as shown in Expression 23 to obtain the in-phase component sampled signal $I1(mT_s)$ and the quadrature component sampled signal $Q1(mT_s)$.

$$\theta = \frac{1}{2^{N-1}} \pi \quad \dots \text{Exp. 23}$$

15 Herein, N is as shown in Table 1. Specifically, where $f_i = 3.072$ [MHz] and $f_s = 24.576$ [MHz], $N = 3$. Where $f_i = 3.072$ [MHz] and $f_s = 12.288$ [MHz], $N = 2$. Where $f_i = 4.608$ [MHz] and $f_s = 36.864$ [MHz], $N = 3$.

For example, a signal $S_b(mT_s)$ whose center frequency is f_b [Hz] being not zero and which has been sampled with the sampling
20 frequency f_s can be converted to a signal whose center frequency is zero using a digital circuit by multiplying $S_b(mT_s)$ by $\exp(-j2\pi \times f_b / f_s \times t)$ and shifting the frequency by f_b in the positive direction. While t represents the time, inside a digital circuit whose sampling frequency is f_s , t cannot take continuous values

but takes discrete values at regular intervals of T_s . Therefore, for every T_s , the value by which $S_b(mT_s)$ should be multiplied is expressed as shown in Expression 24.

$$\exp(-j2\pi \frac{f_b}{f_s} m) \quad \dots \text{Exp. 24}$$

5 Thus, $S_b(mT_s)$ whose center frequency is f_b can be converted to a signal whose center frequency is zero by performing an operation as shown in Expression 25.

$$S_b(mT_s) \times \exp(-j2\pi \frac{f_b}{f_s} m) \quad \dots \text{Exp. 25}$$

Expanding the exp term using the Euler's formula yields
10 Expression 26.

$$\exp(-j2\pi \frac{f_b}{f_s} m) = \cos\{(2\pi \frac{f_b}{f_s})m\} - j\sin\{(2\pi \frac{f_b}{f_s})m\} \quad \dots \text{Exp. 26}$$

Thus, the operation of Expression 25 can be realized by using a circuit configuration as shown in FIG. 9. FIG. 9 is a schematic diagram showing a configuration of a circuit for shifting the
15 frequency by $-f_b$ to obtain a signal whose center frequency is zero. As shown in FIG. 9, the circuit outputs $I_b(mT_s)$ being an in-phase component of $S_b(mT_s)$ and $Q_b(mT_s)$ being a quadrature component thereof. Therefore, the circuit shown in FIG. 9 can be considered a quadrature demodulator. Thus, FIG. 9 shows an internal
20 configuration of the quadrature demodulators 102 and 202.

Therefore, θ in the operation $S_1(mT_s) \times \exp(-j\theta \times m)$ performed

by the quadrature demodulator 102 is determined as shown in Expression 27 below based on Expression 10.

$$\theta = 2\pi \frac{f_b}{f_s} = 2\pi \frac{f_i}{f_s} = 2\pi \frac{f_i}{2^N f_i} = \frac{1}{2^{N-1}} \pi \quad \dots \text{Exp. 27}$$

It can be seen that Expression 27 is equal to Expression 23.

5 Thus, as a result of the operation $S1(mTs) \times \exp(-j\theta \times mTs)$ performed by the quadrature demodulator 102, the in-phase component sampled signal $I1(mTs)$ and the quadrature component sampled signal $Q1(mTs)$ each have a frequency component that is frequency-shifted so that the center frequency of the spectrum 300 shown in FIG.
10 8 is zero.

FIG. 10 is a diagram showing the spectrum of the sampled signal $S2(mTs)$ obtained by sampling the modulated low-frequency signal $L2(t)$ whose center frequency is $f_d = 40.000$ [MHz] with the sampling frequency $f_s = 24.576$ [MHz]. In FIG. 10, the horizontal
15 axis represents the complex frequency. The vertical axis represents the power spectral intensity.

In FIG. 10, a spectrum 500 represents the spectrum of the modulated low-frequency signal $L2(t)$, and the other spectra are folding spectra occurring as a result of sampling the modulated
20 low-frequency signal $L2(t)$ with a sampling period of T_s . Spectra 501, 502 and 503 are each spaced away from the spectrum 500 by a distance of an integer multiple of the sampling frequency, and thus the spectra 500, 501, 502 and 503 are signals equivalent to one another.

However, the other spectra are not spaced apart from the spectrum 500 representing the modulated low-frequency signal $L_2(t)$ by an integer multiple of the sampling frequency, and thus are spectra of signals each having a frequency component different
5 from the spectrum 500.

As described above, the spectrum 500 is the same as the spectrum 502. The quadrature demodulator 202 performs an operation $S_2(mTs) \times \exp(-j\eta \times m)$ using η [rad] expressed as shown in Expression 28 to obtain the in-phase component sampled signal
10 $I_2(mTs)$ and the quadrature component sampled signal $Q_2(mTs)$ each having a frequency component that is frequency-shifted so that the center frequency of the spectrum 502 shown in FIG. 10 is zero.

$$\eta = -\frac{Mf_s - f_d}{f_s} 2\pi \quad \dots \text{Exp. 28}$$

The basis for Expression 28 will now be described. The signal
15 $S_2(mTs)$ outputted from the sampler 201 is equivalent to the spectrum 500, as shown in FIG. 10, and includes a signal whose center frequency is closest to zero. The center frequency of this signal can be expressed as $-Mf_s + f_d$ using a positive integer M . In FIG. 10, $M=2$. It can be seen from Expression 29 that a signal whose center frequency
20 is $-Mf_s + f_d$ can be frequency-shifted to a signal whose center frequency is zero by using η in Expression 28 above.

$$\eta = 2\pi \frac{f_b}{f_s} = 2\pi \frac{-Mf_s + f_d}{f_s} = -\frac{Mf_s - f_d}{f_s} 2\pi \quad \dots \text{Exp. 29}$$

As described above, also where $f_i = 3.072$ [MHz] and $f_s = 12.288$

[MHz], the received data can be properly obtained by performing a similar operation. FIG. 11 is a diagram showing the spectrum of the sampled signal $S1(mTs)$ outputted from the sampler 101 in a case where the center frequency of the modulated low-frequency signal $L1(t)$ is $f_i=3.072$ [MHz] and the sampling frequency f_s is 12.288 [MHz]. In FIG. 11, the horizontal axis represents the complex frequency and the vertical axis represents the power spectral intensity.

In FIG. 11, $2B_{ch}$ represents the channel bandwidth, and $2B_{ch}=5$ [MHz] in the DSRC system. In FIG. 11, a spectrum 400 represents the spectrum of the modulated low-frequency signal $L1(t)$. The other spectra are folding spectra occurring as a result of sampling the modulated low-frequency signal $L1(t)$ with a sampling period of T_s . The figure shows, as folding spectra, a signal whose center frequency is $f_s \pm f_i$ and another signal whose center frequency is $-f_s \pm f_i$.

The quadrature demodulator 102 receives the sampled signal $S1(mTs)$ outputted from the sampler 101, and outputs two signals whose phases are different from each other by $\pi/2$ [rad], i.e., the in-phase component sampled signal $I1(mTs)$ and the quadrature component sampled signal $Q1(mTs)$. The quadrature demodulator 102 can be configured to perform an operation $S1(mTs) \times \exp(-j\theta \times m)$ using θ [rad] expressed as shown in Expression 23, as described above, to obtain the in-phase component sampled signal $I1(mTs)$ and the quadrature component sampled signal $Q1(mTs)$. Also in such a case,

by using the low-pass filter 103 having pass-band characteristics as shown in FIG. 3, it is possible to obtain $Ib1(mTs)$ being an in-phase component signal of the baseband quadrature demodulated signal and $Qb1(mTs)$ being a quadrature component signal thereof
5 each having only a frequency component that is frequency-shifted so that the center frequency of the spectrum 400 is zero. Thus, the received data can be obtained by the received data reproducing section 105.

FIG. 12 is a diagram showing the spectrum of the sampled
10 signal $S2(mTs)$ obtained by sampling the modulated low-frequency signal $L2(t)$ whose center frequency is $f_d=40.000$ [MHz] with the sampling frequency $f_s=12.288$ [MHz]. In FIG. 12, the horizontal axis represents the complex frequency. The vertical axis represents the power spectral intensity.

15 In FIG. 12, a spectrum 704 represents the spectrum of the modulated low-frequency signal $L2(t)$, and the other spectra are folding spectra occurring as a result of sampling the modulated low-frequency signal $L2(t)$ with a sampling period of T_s . Spectra
705, 706 and 707 are each spaced away from the spectrum 704 by
20 a distance of an integer multiple of the sampling frequency, and thus the spectra 704, 705, 706 and 707 are signals equivalent to one another.

However, the other spectra are not spaced apart from the spectrum 704 representing the modulated low-frequency signal $L2(t)$
25 by an integer multiple of the sampling frequency, and thus are

spectra of signals each having a frequency component different from the spectrum 704.

As described above, the spectrum 704 is the same as the spectrum 707. The quadrature demodulator 202 performs an operation $S2(mTs) \times \exp(-j\eta \times m)$ using η [rad] expressed as shown in Expression 28 to obtain the in-phase component sampled signal $I2(mTs)$ and the quadrature component sampled signal $Q2(mTs)$ each having a frequency component that is frequency-shifted so that the center frequency of the spectrum 707 shown in FIG. 12 is zero. In this case, since the spectrum 707 is frequency-shifted so that the center frequency thereof is zero, $M=3$ in Expression 28.

Also where $f_i=4.608$ [MHz] and $f_s=36.864$ [MHz], the received data can be properly obtained at the first and second wireless digital receivers 21 and 31 similarly by determining the rotation angles θ and η at the quadrature demodulators 102 and 202 and performing the quadrature demodulation.

Thus, in the first embodiment, the sampling frequency in the base station and that in the mobile station are both set to the same value f_s [Hz] being an even-number multiple of the wireless symbol transmission rate such that oversampling is done in the base station while undersampling is done in the mobile station. Moreover, the center frequency f_i [Hz] of a signal that has been downconverted in the base station is $1/2$ to 1 times the frequency corresponding to the bandwidth of the transmitted/received wireless signal and is $1/2^N$ (N is a natural number) times the sampling

frequency. For example, $f_i=3.072$ [MHz] and $f_s=24.576$ [MHz]. Alternatively, $f_i=3.072$ [MHz] and $f_s=12.288$ [MHz]. Alternatively, $f_i=4.608$ [MHz] and $f_s=36.864$ [MHz], $f_i=3.584$ [MHz] and $f_s=28.672$ [MHz], or $f_i=4.096$ [MHz] and $f_s=32.768$ [MHz]. Thus, the
5 demodulation digital circuit in the base station and that in the mobile station can be the same except for using different rotation angles in the quadrature demodulation. Therefore, it is possible to provide wireless digital receivers for the base station and for the mobile station at a low cost while reducing the overall
10 cost of the wireless communications system.

Note that in FIG. 2 being a block diagram showing a functional configuration of the first wireless digital receiver 21, the frequency converter 100, the sampler 101, the quadrature demodulator 102, the low-pass filter 103, the sampling signal
15 generator 104 and the received data reproducing section 105 are typically each implemented in the form of an LSI being an integrated circuit. These components may be individually formed into a separate chip, or some or all of them may be formed together into
a single chip.

20 Moreover, in FIG. 4 being a block diagram showing a functional configuration of the second wireless digital receiver 31, the frequency converter 200, the sampler 201, the quadrature demodulator 202, the low-pass filter 203, the sampling signal generator 204 and the received data reproducing section 205 are
25 typically each implemented in the form of an LSI being an integrated

circuit. These components may be individually formed into a separate chip, or some or all of them may be formed together into a single chip.

While the term "LSI" is used herein as the type of integrated circuit used in the present invention, integrated circuits are also called "ICs", "system LSIs", "super LSIs" or "ultra LSIs" depending on the degree of integration. Moreover, the form of an integrated circuit that can be used with the present invention is not limited to an LSI, but may alternatively be a dedicated circuit or a general-purpose processor. It may alternatively be an FPGA (Field Programmable Gate Array) being programmable after the LSI is manufactured, or a reconfigurable processor in which the interconnections and settings of circuit cells in the LSI can be reconfigured. Furthermore, if advancements in the semiconductor technology or derivative technologies bring forth a new form of circuit integration replacing LSIs, the new form of circuit integration can of course be used for the integration of the frequency converter 100, the sampler 101, the quadrature demodulator 102, the low-pass filter 103, the sampling signal generator 104 and the received data reproducing section 105.

Similarly, such a new form of circuit integration can be used for the integration of the frequency converter 200, the sampler 201, the quadrature demodulator 202, the low-pass filter 203, the sampling signal generator 204 and the received data reproducing section 205.

Such a derivative technology may possibly be an application of biotechnology, for example.

(Second Embodiment)

In the first embodiment, a modulated low-frequency signal
5 obtained by converting the frequency of a modulated high-frequency
signal is sampled, and then an in-phase component sampled signal
and a quadrature component sampled signal whose phases are
different from each other by $\pi/2$ are outputted by the quadrature
demodulator, which are then low-pass filtered using a low-pass
10 filter, thereby obtaining the received data. A second embodiment
of the present invention is directed to a wireless digital receiver
using a complex filter instead of the quadrature demodulator and
the low-pass filter to obtain received data. The overall system
configuration of the second embodiment is similar to that of the
15 first embodiment, and thus FIG. 1 will be relied upon also in the
second embodiment.

FIG. 13 is a block diagram showing a functional configuration
of the first wireless digital receiver 21 according to the second
embodiment of the present invention. In FIG. 13, the first wireless
20 digital receiver 21 includes a frequency converter 600, a sampler
601, the complex filter 602, a sampling signal generator 603 and
a received data reproducing section 604. The complex filter 602
and the received data reproducing section 604 will be hereinafter
referred to collectively as a "demodulation digital circuit".

25 In the first wireless digital receiver 21, the frequency

converter 600 converts the frequency of the modulated high-frequency signal $R(t)$ into a modulated low-frequency signal $L(t)$ whose center frequency is f_i . The sampler 601 samples the modulated low-frequency signal $L(t)$ with a sampling signal whose
5 sampling frequency is f_s outputted from the sampling signal generator 603 to output a sampled signal $S(mT_s)$. The operation hitherto is similar to that of the first embodiment.

Therefore, the spectrum of the sampled signal $S(mT_s)$ outputted from the sampler 601 is the same as that in a case where
10 the center frequency of the modulated low-frequency signal $L(t)$ is $f_i=3.072$ [MHz] and the sampling frequency is $f_s=24.576$ [MHz], i.e., that shown in FIG. 8. Therefore, FIG. 8 will be relied upon also in the second embodiment.

In FIG. 8, spectra equivalent to the spectrum 300, being
15 the spectrum of the modulated low-frequency signal $L(t)$, are those spaced apart from the spectrum 300 by an integer multiple of the sampling frequency $f_s=24.576$ [MHz]. Thus, the spectrum 300 and the spectrum whose center frequency is -3.072 [MHz] are spectra having different characteristics. In order to obtain received
20 data, the spectrum 300 or a spectrum having a spectrum spaced apart from the spectrum 300 by an integer multiple of the sampling frequency f_s as a frequency component should be extracted. FIG. 14 is a diagram showing exemplary pass-band characteristics of the complex filter 602. Where the complex filter 602 having
25 pass-band characteristics as shown in FIG. 14 is used, $I_b(mT_s)$

being an in-phase component of the quadrature demodulated signal outputted from the complex filter 602 and $Q_b(mTs)$ being a quadrature component thereof are signals having the spectrum 300 as a frequency component and whose phases are different from each other by $\pi/2$ 5 [rad]. Although the quadrature demodulated signals $I_b(mTs)$ and $Q_b(mTs)$ are signals whose center frequencies are not zero, the received data reproducing section 604 can output the received data by means of delay detection, or the like.

In the second embodiment, the configuration of the second 10 wireless digital receiver 31 in the mobile station is similar to the configuration of the first wireless digital receiver 21, and thus FIG. 13 will be relied upon also for the configuration of the second wireless digital receiver 31.

The second wireless digital receiver 31 is different from 15 the first wireless digital receiver 21 in that the modulated high-frequency signal $R(t)$ is converted by the frequency converter 600 to the modulated low-frequency signal $L(t)$ whose center frequency is f_d and in that a filter that extracts the spectrum 502 whose center frequency is -9.152 [MHz] as shown in FIG. 10 20 is used as the complex filter. Otherwise, the first wireless digital receiver 21 is the same as the second wireless digital receiver 31.

Thus, in the second embodiment, the received data can be obtained only by changing the pass-band characteristics of the 25 complex filter, thus obtaining effects similar to those of the

first embodiment.

The above description is directed to a case where $f_i=3.072$ [MHz] and $f_s=24.576$ [MHz]. Also where $f_i=3.072$ [MHz] and $f_s=12.288$ [MHz], similar results can be obtained only by changing the center frequency of the pass-band characteristics of the complex filter 602 from 3.072 [MHz] to 3.136 [MHz] (see FIG. 12). Also where $f_i=3.584$ [MHz] and $f_s=28.672$ [MHz], similar results can be obtained only by changing the center frequency of the pass-band characteristics of the complex filter 602 from 3.072 [MHz] to 3.584 [MHz] (see FIG. 15A). Also where $f_i=4.096$ [MHz] and $f_s=32.768$ [MHz], similar results can be obtained only by changing the center frequency of the pass-band characteristics of the complex filter 602 from 3.072 [MHz] to 4.096 [MHz] (see FIG. 15B). Also where $f_i=4.608$ [MHz] and $f_s=36.864$ [MHz], similar results can be obtained only by changing the center frequency of the pass-band characteristics of the complex filter 602 from 3.072 [MHz] to 4.608 [MHz] (see FIG. 15C).

Also where the center frequency of the modulated low-frequency signal $L(t)$ inputted to the sampler 601 is $f_d=40.000$ [MHz] and $f_s=24.576$ [MHz] or $f_s=12.288$ [MHz], similar results can be obtained only by changing the center frequency of the pass-band characteristics of the complex filter 602 to -9.152 [MHz] or 3.136 [MHz], respectively. Also where $f_d=40.000$ [MHz] and $f_s=28.672$ [MHz], similar results can be obtained only by changing the center frequency of the pass-band characteristics of the complex filter

602 to 11.328 [MHz] (see FIG. 15D). Also where $f_d=40.000$ [MHz] and $f_s=32.768$ [MHz], similar results can be obtained only by changing the center frequency of the pass-band characteristics of the complex filter 602 to 7.232 [MHz] (see FIG. 15E). Also
5 where $f_d=40.000$ [MHz] and $f_s=36.864$ [MHz], similar results can be obtained only by changing the center frequency of the pass-band characteristics of the complex filter 602 to 3.136 [MHz] (see FIG. 15F).

As for the complex filter characteristics, in a case where
10 an FIR (Finite Impulse Response) filter is used as the complex filter, for example, the number of taps can be determined in advance so as to accommodate any of the center frequencies of the pass-band characteristics of 3.072 [MHz], 3.136 [MHz] and -9.152 [MHz], whereby all of the cases mentioned above can be addressed only
15 by selecting an appropriate tap coefficient. Thus, by using an FIR with which one of different tap coefficients can be selected, the same demodulation digital circuit, being a complex filter, can be used for the mobile station and for the base station, whereby
it is possible to reduce the cost.

20 Note that in FIG. 13 being a block diagram showing a functional configuration of the first wireless digital receiver 21 according to the second embodiment of the present invention, the frequency converter 600, the sampler 601, the complex filter 602, the sampling signal generator 603 and the received data reproducing section
25 604 are typically each implemented in the form of an LSI being

an integrated circuit. These components may be individually formed into a separate chip, or some or all of them may be formed together into a single chip.

While the term "LSI" is used herein as the type of integrated circuit used in the present invention, integrated circuits are also called "ICs", "system LSIs", "super LSIs" or "ultra LSIs" depending on the degree of integration. Moreover, the form of an integrated circuit that can be used with the present invention is not limited to an LSI, but may alternatively be a dedicated circuit or a general-purpose processor. It may alternatively be an FPGA (Field Programmable Gate Array) being programmable after the LSI is manufactured, or a reconfigurable processor in which the interconnections and settings of circuit cells in the LSI can be reconfigured. Furthermore, if advancements in the semiconductor technology or derivative technologies bring forth a new form of circuit integration replacing LSIs, the new form of circuit integration can of course be used for the integration of the frequency converter 600, the sampler 601, the complex filter 602, the sampling signal generator 603 and the received data reproducing section 604. Such a derivative technology may possibly be an application of biotechnology, for example.

(Third Embodiment)

FIG. 1 will be relied upon also in a third embodiment of the present invention. FIG. 16A and FIG. 16B are block diagrams each showing a functional configuration of the first wireless

digital receiver 21 according to the third embodiment of the present invention.

In FIG. 16A, the first wireless digital receiver 21 includes a frequency converter 800, a sampler 801, the quadrature demodulator 802, an automatic frequency controller 803, a low-pass filter 804, a sampling signal generator 805, a detector 806 and a data determination section 807. The quadrature demodulator 802, the automatic frequency controller 803, the low-pass filter 804, the detector 806 and the data determination section 807 will be hereinafter referred to collectively as a "demodulation digital circuit".

In the third embodiment, the frequency converter 800 does not convert the modulated high-frequency signal $R(t)$ to a modulated low-frequency signal whose center frequency is 3.072 [MHz]. The following description is directed to a case where the frequency converter 800 converts the modulated high-frequency signal $R(t)$ to the modulated low-frequency signal $L(t)$ whose center frequency is $f_j=3.000$ [MHz].

The sampler 801 samples the modulated low-frequency signal $L(t)$ in synchronism with the sampling signal whose frequency is $f_s=24.576$ [MHz] outputted from the sampling signal generator 805 to output the sampled signal $S(mTs)$.

The quadrature demodulator 802 assumes that the center frequency of the modulated low-frequency signal $L(t)$ is $f_i=3.072$ [MHz], and performs an operation $S(mTs) \times \exp(-j\theta \times mTs)$ using θ [rad]

expressed as shown in Expression 23 to obtain the in-phase component sampled signal $I(mTs)$ and the quadrature component sampled signal $Q(mTs)$.

FIG. 17 is a diagram showing the spectrum of the in-phase component sampled signal $I(mTs)$ and the quadrature component sampled signal $Q(mTs)$ outputted from the quadrature demodulator 802. In FIG. 17, a spectrum 900 is the spectrum of the modulated low-frequency signal $L(t)$. The other spectra are folding spectra occurring as a result of sampling the modulated low-frequency signal $L(t)$ with a sampling period of T_s . A comparison between FIG. 17 and FIG. 8 shows that these spectra as a whole are shifted from each other by 0.072 [MHz], which is the difference between 3.072 [MHz] being the intended frequency of the modulated low-frequency signal $L(t)$ and the actual frequency 3.000 [MHz] thereof.

The automatic frequency controller 803 converts the frequency of the spectrum 900 so that it is frequency-shifted to its intended center frequency of 3.072 [MHz]. In other words, the automatic frequency controller 803 converts the entire spectrum shown in FIG. 17 so that the center frequency of the spectrum 900 is 3.072 [MHz]. Such an automatic frequency controller 803 is disclosed in Japanese Patent No. 3327152, Japanese Laid-Open Patent Publication No. 6-120997, etc.

If the automatic frequency controller 803 performing such an operation is provided between the quadrature demodulator 802

and the low-pass filter 804, the low-pass filter 804 can be a filter having the same pass-band characteristics as those shown in FIG.

3. The detector 806 provided on the output side of the low-pass filter 804 performs a delay detection operation to output detection
5 signals DETI(mTs) and DETQ(mTs) to the data determination section 807. The data determination section 807 detects a phase using the signals DETI(mTs) and DETQ(mTs), and outputs the received data based on the detected phase.

Thus, the third embodiment provides the following advantage.

10 In a case where f_i as calculated in the first embodiment cannot be used, e.g., where it is necessary to order a tailored frequency oscillator in order to use f_i as calculated in the first embodiment, it is possible to obtain a sampled signal having a component whose center frequency is f_i by digitally correcting the frequency with
15 an automatic frequency controller by using a frequency converter capable of converting a frequency to another frequency near f_i . Thus, the received data can be properly reproduced. By providing a frequency converter using a general-purpose local oscillator so that a frequency can be converted to another frequency near
20 f_i , it is possible to reduce the cost of the wireless digital receiver.

While the above description is directed to a case where the automatic frequency controller 803 is provided immediately after the quadrature demodulator 802, similar effects can be obtained
25 also with a configuration as shown in FIG. 16B. Note however that

where a configuration as shown in FIG. 16B is used, it is necessary to use an automatic frequency controller as disclosed in Japanese Patent No. 3088893, Japanese Laid-Open Patent Publication No. 10-98500, etc.

5 While the above description is directed to a case where the center frequency of the modulated low-frequency signal $L(t)$ is shifted from f_i , similar effects can be obtained also where the center frequency of the modulated low-frequency signal $L(t)$ is shifted from f_d . Specifically, by performing the frequency
10 shifting operation at the automatic frequency controller 803 so that the center frequency of the spectrum of the modulated low-frequency signal $L(t)$ is equal to f_d , pass-band characteristics as shown in FIG. 3 can be used as the pass-band characteristics of the low-pass filter 804, and the received data can be obtained
15 by a delay detection circuit, or the like, provided on the output side of the low-pass filter 804.

While $f_j = 3.000$ [MHz] in the above description, the present invention is not limited to this as long as the frequency shift Δf between $f_i = 3.072$ and f_j satisfies $|\Delta f| < 0.512$ [MHz]. The reason
20 for this will now be described. A frequency is an amount of phase change per unit time. Therefore, there is a one-to-one correspondence between a frequency shift and a phase shift. The DSRC system uses a format of transmitted data in which the beginning portion of each frame contains a preamble pattern made up of symbols
25 each having a phase different from that of the next symbol by π .

By using the preamble pattern, a phase correction of up to $\pm\pi/2$ (excluding $\pm\pi/2$) can be performed in principle. Where a phase difference of $\pi/2$ is converted to a frequency, the symbol data rate f_{sym} is involved in the conversion formula, whereby the
 5 frequency for $\pi/2$ varies depending on the value of f_{sym} . This is expressed in Expression 30.

$$\theta_{\text{err}} = 2\pi \times \frac{\Delta f}{f_s} \text{ [rad]} \quad \dots \text{Exp. 30}$$

Herein, θ_{err} is the phase for the frequency shift Δf . In the present embodiment, $f_{\text{sym}} = 2.048$ [MHz]. Therefore, where $\theta_{\text{err}} = \pi/2$ and
 10 $f_s = 2.048$ [MHz], Expression 30 can be rearranged with respect to Δf to yield $|\Delta f| < 0.512$ [MHz].

The above description is directed to a case where the automatic frequency controller 803 being a circuit for correcting a signal whose center frequency is f_j to a signal whose center
 15 frequency is f_i is provided following the sampler 801. Alternatively, such a frequency correction circuit for correcting a frequency may be provided preceding the sampler 801. Thus, the low-frequency signal downconverted by the frequency converter 800 may be demodulated after being corrected to a signal whose center
 20 frequency is f_i at a position either preceding or following the sampler 801.

Note that in FIG. 16A being a block diagram showing a functional configuration of the first wireless digital receiver
 21 according to the third embodiment of the present invention,

the frequency converter 800, the sampler 801, the quadrature demodulator 802, the automatic frequency controller 803, the low-pass filter 804, the sampling signal generator 805, the detector 806 and the data determination section 807 are typically
5 each implemented in the form of an LSI being an integrated circuit. These components may be individually formed into a separate chip, or some or all of them may be formed together into a single chip.

Also where a configuration as shown in FIG. 16B is used, the frequency converter 800, the sampler 801, the quadrature
10 demodulator 802, the automatic frequency controller 803, the low-pass filter 804, the sampling signal generator 805, the detector 806 and the data determination section 807 may be individually formed into a separate chip, or some or all of them may be formed together into a single chip.

15 While the term "LSI" is used herein as the type of integrated circuit used in the present invention, integrated circuits are also called "ICs", "system LSIs", "super LSIs" or "ultra LSIs" depending on the degree of integration. Moreover, the form of
an integrated circuit that can be used with the present invention
20 is not limited to an LSI, but may alternatively be a dedicated circuit or a general-purpose processor. It may alternatively be an FPGA (Field Programmable Gate Array) being programmable after the LSI is manufactured, or a reconfigurable processor in which the interconnections and settings of circuit cells in the LSI can
25 be reconfigured. Furthermore, if advancements in the

semiconductor technology or derivative technologies bring forth a new form of circuit integration replacing LSIs, the new form of circuit integration can of course be used for the integration of the frequency converter 800, the sampler 801, the quadrature
5 demodulator 802, the automatic frequency controller 803, the low-pass filter 804, the sampling signal generator 805, the detector 806 and the data determination section 807.

Also where a configuration as shown in FIG. 16B is used, such a new form of circuit integration replacing LSIs brought forth
10 by advancements in the semiconductor technology or derivative technologies may be used for the integration of the frequency converter 800, the sampler 801, the quadrature demodulator 802, the automatic frequency controller 803, the low-pass filter 804, the sampling signal generator 805, the detector 806 and the data
15 determination section 807.

Such a derivative technology may possibly be an application of biotechnology, for example.

(Fourth Embodiment)

A fourth embodiment of the present invention is directed
20 to a base-station wireless communications device obtained by combining together the first wireless transmitter and the first wireless digital receiver in the base station, and a mobile-station wireless communications device obtained by combining together the
second wireless transmitter and the second wireless digital
25 receiver in the mobile station.

FIG. 18 is a diagram showing a configuration of a base-station wireless communications device 12 according to the fourth embodiment of the present invention. In FIG. 18, the base-station wireless communications device 12 includes an antenna 1200, a band-pass filter 1216, a transmission/reception selector switch 1211, an amplifier 1201, a first mixer 1202, a second mixer 1203, a first local oscillator 1206, a first low-pass filter 1204, a second low-pass filter 1205, a first sampler 1207, a second sampler 1208, a sampling signal generator 1209, a demodulation digital circuit 1210, a transmission high-frequency circuit 1212, a third mixer 1213, a second local oscillator 1214 and a transmitter circuit 1215.

In the base-station wireless communications device 12, the signal-receiving operation is performed by using the antenna 1200, the band-pass filter 1216, the transmission/reception selector switch 1211, the amplifier 1201, the first mixer 1202, the second mixer 1203, the first local oscillator 1206, the first low-pass filter 1204, the second low-pass filter 1205, the first sampler 1207, the second sampler 1208, the sampling signal generator 1209 and the demodulation digital circuit 1210. The signal-transmitting operation is performed by using the transmitter circuit 1215, the second local oscillator 1214, the third mixer 1213, the transmission high-frequency circuit 1212, the transmission/reception selector switch 1211, the band-pass filter 1216 and the antenna 1200.

In the signal-receiving operation, the transmission/reception selector switch 1211 is switched so that the antenna 1200 and the amplifier 1201 are connected to each other. The modulated high-frequency signal $R(t)$ received by the antenna
5 1200 from the mobile station whose center frequency is f_c is first passed through the band-pass filter 1216 to remove signals of frequency bands that are used neither in the base station nor in the mobile station, and is then inputted to the amplifier 1201. The amplifier 1201 amplifies the modulated high-frequency signal
10 $R(t)$ to an appropriate level, and inputs the amplified signal to the first mixer 1202 and the second mixer 1203. The first local oscillator 1206 outputs a sine wave whose center frequency is $f_c - f_i$. Herein, f_i is 3.072 [MHz] as calculated in the first embodiment.

The first mixer 1202 multiplies the sine wave outputted from
15 the first local oscillator 1206 whose center frequency is $f_c - f_i$ with the modulated high-frequency signal $R(t)$ to output a modulated low-to-intermediate-frequency signal in-phase component $R_{XI}(t)$ whose center frequency is f_i . The first low-pass filter 1204
removes a high-frequency component from the modulated
20 low-to-intermediate-frequency signal in-phase component $R_{XI}(t)$, and passes the filtered signal to the first sampler 1207.

The second mixer 1203 multiplies a signal outputted from the first local oscillator 1206 whose center frequency is $f_c - f_i$ and whose phase is shifted from that of the sine wave by $\pi/2$ with
25 the modulated high-frequency signal $R(t)$ to output a modulated

low-to-intermediate-frequency signal quadrature component $RXQ(t)$ whose center frequency is f_i . The second low-pass filter 1205 removes a high-frequency component from the modulated low-to-intermediate-frequency signal quadrature component $RXQ(t)$,
5 and passes the filtered signal to the second sampler 1208.

The first sampler 1207 samples the modulated low-to-intermediate-frequency signal in-phase component $RXI(t)$ in synchronism with a signal outputted from the sampling signal generator 1209 whose frequency is $f_s=24.576$ [MHz] to output the
10 in-phase component sampled signal $I(mTs)$.

The second sampler 1208 samples the modulated low-to-intermediate-frequency signal quadrature component $RXQ(t)$ in synchronism with a signal outputted from the sampling signal generator 1209 whose frequency is $f_s=24.576$ [MHz] to output the
15 quadrature component sampled signal $Q(mTs)$.

The demodulation digital circuit 1210 receives the in-phase component sampled signal $I(mTs)$ and the quadrature component sampled signal $Q(mTs)$, and performs a quadrature demodulation operation on the received signals. Then, the demodulation digital
20 circuit 1210 low-pass-filters the demodulated signals to output received data.

In FIG. 18, the first and second mixers 1202 and 1203, the first local oscillator 1206 and the first and second low-pass filters 1204 and 1205 correspond to the frequency converter 100
25 in the first embodiment. The first and second samplers 1207 and

1208 correspond to the sampler 101 illustrated in the first embodiment. In the fourth embodiment, quadrature data is sampled, unlike in the first embodiment. However, the fourth embodiment is substantially the same as the first embodiment since the values
5 of f_i and f_s used in the first embodiment are used also in the fourth embodiment. The sampling signal generator 1209 corresponds to the sampling signal generator 104 illustrated in the first embodiment. The demodulation digital circuit 1210 corresponds to the quadrature demodulator 102, the low-pass filter
10 103 and the received data reproducing section 105 in the first embodiment.

In the signal-transmitting operation, data to be transmitted is modulated according to the $\pi/4$ shift QPSK scheme in the transmitter circuit 1215, and is outputted as a transmitted signal
15 $B(t)$. The third mixer 1213 multiplies the transmitted signal $B(t)$ by a signal outputted from the local oscillator 1214 whose center frequency is $f_c - f_d$ to output a modulated high-frequency signal $TX(t)$. The modulated high-frequency signal $TX(t)$ is passed through the transmission high-frequency circuit 1212 to remove
20 unnecessary frequency components, and adjusted to an appropriate transmission power level, after which the signal is radiated off the antenna 1200 in the form of a radio wave.

FIG. 19 is a diagram showing a configuration of a mobile-station wireless communications device 11 according to the
25 fourth embodiment of the present invention. In FIG. 19, the

mobile-station wireless communications device 11 includes an antenna 1100, a band-pass filter 1112, a transmission/reception selector switch 1108, an amplifier 1101, a first mixer 1102, a local oscillator 1103, a low-pass filter 1104, a sampler 1105, 5 a sampling signal generator 1106, a demodulation digital circuit 1107, a transmission high-frequency circuit 1109, a second mixer 1110 and a transmitter circuit 1111.

In the mobile-station wireless communications device, the signal-receiving operation is performed by using the antenna 1100, 10 the band-pass filter 1112, the transmission/reception selector switch 1108, the amplifier 1101, the first mixer 1102, the local oscillator 1103, the low-pass filter 1104, the sampler 1105, the sampling signal generator 1106 and the demodulation digital circuit 1107. The signal-transmitting operation is performed by using 15 the transmitter circuit 1111, the second mixer 1110, the local oscillator 1103, the transmission high-frequency circuit 1109, the transmission/reception selector switch 1108, the band-pass filter 1112 and the antenna 1100.

In the signal-receiving operation, the 20 transmission/reception selector switch 1108 is switched so that the antenna 1100 and the amplifier 1101 are connected to each other. The modulated high-frequency signal $RL(t)$ from the base station received by the antenna 1100 whose center frequency is $f_c - f_d$ is first passed through the band-pass filter 1112 to remove signals 25 of frequency bands that are used neither in the base station nor

in the mobile station, and is then inputted to the amplifier 1101. The amplifier 1101 amplifies the modulated high-frequency signal $RL(t)$ to an appropriate level, and inputs the amplified signal to the first mixer 1102. The first local oscillator 1103 outputs
5 a sine wave whose center frequency is f_c .

The first mixer 1102 multiplies the sine wave outputted from the local oscillator 1103 whose center frequency is f_c with the modulated high-frequency signal $RL(t)$ to output a modulated low-to-intermediate-frequency signal $L(t)$ whose center frequency
10 is f_d to the low-pass filter 1104. In the DSRC system, the frequency difference between the downlink and the uplink is 40.000 [MHz]. Therefore, $f_d=40.000$ [MHz]. The low-pass filter 1104 removes a high-frequency component from the modulated low-to-intermediate-frequency signal $L(t)$, and passes the
15 filtered signal to the sampler 1105.

The sampler 1105 samples the modulated low-to-intermediate-frequency signal $L(t)$ in synchronism with a signal outputted from the sampling signal generator 1106 whose frequency is $f_s=24.576$ [MHz] to output a sampled signal $L_s(mT_s)$.

20 The demodulation digital circuit 1107 receives the sampled signal $L_s(mT_s)$, and performs a quadrature demodulation operation on the received signal. Then, the demodulation digital circuit 1107 low-pass-filters the demodulated signal to output received data.

25 In FIG. 19, the first mixer 1102, the local oscillator 1103

and the low-pass filter 1104 correspond to the frequency converter 200 in the first embodiment. The sampler 1105 corresponds to the sampler 201 in the first embodiment. The sampling signal generator 1106 corresponds to the sampling signal generator 204 in the first
5 embodiment. The demodulation digital circuit 1107 corresponds to the quadrature demodulator 202, the low-pass filter 203 and the received data reproducing section 205 in the first embodiment.

In the signal-transmitting operation, data to be transmitted is modulated according to the $\pi/4$ shift QPSK scheme in the
10 transmitter circuit 1111, and is outputted as the transmitted signal $B(t)$. The second mixer 1110 multiplies the transmitted signal $B(t)$ by a signal outputted from the local oscillator 1103 whose center frequency is f_c to output a modulated high-frequency signal $TX(t)$. The modulated high-frequency signal $TX(t)$ is passed
15 through the transmission high-frequency circuit 1109 to remove unnecessary frequency components, and adjusted to an appropriate transmission power level, after which the signal is radiated off the antenna 1100 in the form of a radio wave.

Thus, in the fourth embodiment, the same sampling frequency
20 is used for the mobile station and for the base station, and thus the same demodulation digital circuit can be used for the mobile station and for the base station, whereby it is possible to provide a wireless communications system and a wireless digital receiver for use therein at a low cost.

25 Note that the components of the base-station wireless

communications device 12 shown in FIG. 18, i.e., the antenna 1200, the band-pass filter 1216, the transmission/reception selector switch 1211, the amplifier 1201, the first mixer 1202, the second mixer 1203, the first local oscillator 1206, the first low-pass
5 filter 1204, the second low-pass filter 1205, the first sampler 1207, the second sampler 1208, the sampling signal generator 1209, the demodulation digital circuit 1210, the transmission high-frequency circuit 1212, the third mixer 1213, the second local oscillator 1214 and the transmitter circuit 1215, are typically
10 each implemented in the form of an LSI being an integrated circuit. These components may be individually formed into a separate chip, or some or all of them may be formed together into a single chip.

Similarly, the components of the mobile-station wireless communications device 11 shown in FIG. 19, i.e., the antenna 1100,
15 the band-pass filter 1112, the transmission/reception selector switch 1108, the amplifier 1101, the first mixer 1102, the local oscillator 1103, the low-pass filter 1104, the sampler 1105, the sampling signal generator 1106, the demodulation digital circuit 1107, the transmission high-frequency circuit 1109, the second
20 mixer 1110 and the transmitter circuit 1111, are typically each implemented in the form of an LSI being an integrated circuit. These components may be individually formed into a separate chip, or some or all of them may be formed together into a single chip.

While the term "LSI" is used herein as the type of integrated
25 circuit used in the present invention, integrated circuits are

also called "ICs", "system LSIs", "super LSIs" or "ultra LSIs" depending on the degree of integration. Moreover, the form of an integrated circuit that can be used with the present invention is not limited to an LSI, but may alternatively be a dedicated
5 circuit or a general-purpose processor. It may alternatively be an FPGA (Field Programmable Gate Array) being programmable after the LSI is manufactured, or a reconfigurable processor in which the interconnections and settings of circuit cells in the LSI can be reconfigured. Furthermore, if advancements in the
10 semiconductor technology or derivative technologies bring forth a new form of circuit integration replacing LSIs, the new form of circuit integration can of course be used for the integration of the antenna 1200, the band-pass filter 1216, the transmission/reception selector switch 1211, the amplifier 1201,
15 the first mixer 1202, the second mixer 1203, the first local oscillator 1206, the first low-pass filter 1204, the second low-pass filter 1205, the first sampler 1207, the second sampler 1208, the sampling signal generator 1209, the demodulation digital
circuit 1210, the transmission high-frequency circuit 1212, the
20 third mixer 1213, the second local oscillator 1214 and the transmitter circuit 1215.

Similarly, such a new form of circuit integration replacing LSIs brought forth by advancements in the semiconductor technology or derivative technologies may be used for the integration of the
25 antenna 1100, the band-pass filter 1112, the

transmission/reception selector switch 1108, the amplifier 1101, the first mixer 1102, the local oscillator 1103, the low-pass filter 1104, the sampler 1105, the sampling signal generator 1106, the demodulation digital circuit 1107, the transmission
5 high-frequency circuit 1109, the second mixer 1110 and the transmitter circuit 1111.

Such a derivative technology may possibly be an application of biotechnology, for example.

While the DSRC system has been described in detail in the
10 first to fourth embodiments, it is understood that a wireless communications system and a wireless data receiver providing similar effects can be obtained with other types of FDD systems.

The various functional blocks mentioned above in the first to fourth embodiments may be any means capable of performing their
15 functions. For example, the frequency converter may be any frequency converting means, the sampler may be any sampling means, the demodulation digital circuit may be any digital demodulating means, the quadrature demodulator may be any quadrature demodulating means, the low-pass filter may be any low-pass
20 filtering means, the received data reproducing section may be any received data reproducing means, and the complex filter may be any complex filtering means. These functional blocks are not limited to any particular types of devices as long as they are operable to perform their functions.

25 While the invention has been described in detail, the

foregoing description is in all aspects illustrative and not restrictive. It is understood that numerous other modifications and variations can be devised without departing from the scope of the invention.

5

INDUSTRIAL APPLICABILITY

A wireless communications system and a wireless digital receiver for use therein according to the present invention can be provided at a low cost, and are useful in various applications
10 such as a wireless communications application using an FDD architecture.